Proceedings of the I·R·E

JULY 1941

VOLUME 29 NUMBER 7

Institute of Radio Engineers

Proceedings of the I·R·E

Published Monthly by
The Institute of Radio Engineers, Inc.

VOLUME 29 — *July, 1941* — NUMBER 7

Entered as second-class matter October 26, 1927, at the post office at Menasha, Wisconsin, under the Act of February 28, 1925, embodied in Paragraph 4, Section 538 of the Postal Laws and Regulations. Publication office, 450 Ahnaip Street, Menasha, Wisconsin. Editorial and advertising offices, 330 West 42nd St., New York, N. Y. Subscription, $10.00 per year; foreign, $11.00.

THE INSTITUTE OF RADIO ENGINEERS
INCORPORATED

Pacific Coast Convention—Seattle, Washington—August 20, 21, and 22, 1941

Rochester Fall Meeting, November 10, 11, and 12, 1941

SECTIONS

ATLANTA—Chairman, A. W. Shropshire, Secretary, G. M. Howard 856 St. Charles Ave., N. E., Atlanta, Ga.

BALTIMORE—Chairman, Ferdinand Hamburger, Jr.; Secretary, G. J. Gross, Pennsylvania Water and Power Co., 1611 Lexington Bldg., Baltimore, Md.

BOSTON—Chairman, J. M. Henry; Secretary, R. O. Oberg, Northeastern University, Boston, Mass.

BUENOS AIRES—Chairman, P. J. Noizeux; Secretary, L. C. Simpson, RCA Victor Argentina, Bartolome Mitre, 1961, Buenos Aires, Argentina.

BUFFALO-NIAGARA—Chairman, E. H. Roy; Secretary, Leroy Fiedler, 53 Rosedale Ave., Hamburg, N. Y.

CHICAGO—Chairman, G. I. Martin; Secretary, R. A. Kay, RCA Institutes, Inc., 1154 Merchandise Mart, Chicago, Ill.

CINCINNATI—Chairman, J. M. McDonald; Secretary, W. L. Schwesinger, Crosley Corp., Cincinnati, Ohio.

CLEVELAND—Chairman, Carl E. Smith; Secretary, W. G. Hutton, 7314 Dorothy Ave., Parma, Ohio.

CONNECTICUT VALLEY—Chairman, K. A. McLeod; Secretary, W. M. Smith, 250 Main St., Suffield, Conn.

DALLAS-FORT WORTH—Chairman, D. A. Peterson; Secretary, P. C. Barnes, WFAA-WBAP, Grapevine, Tex.

DETROIT—Chairman, M. Cottrell; Secretary, Paul Frincke, WJBK, 6559 Hamilton Ave., Detroit, Mich.

EMPORIUM—Chairman, R. K. Gessford; Secretary, C. W. Reash, Hygrade Sylvania Corp., Emporium, Pa.

INDIANAPOLIS—Chairman, Arthur Curtiss; Secretary, T. N. Rosser, P. R. Mallory & Co., E. Washington St., Indianapolis, Ind.

KANSAS CITY—Chairman, Harner Selvidge; Secretary, G. L. Taylor, Midland Radio School, Power and Light Bldg., Kansas City, Mo.

LOS ANGELES—Chairman, W. W. Lindsay, Jr.; Secretary, C. F. Wolcott, 1609 S. Western Ave., Los Angeles, Calif.

MONTREAL—Chairman, E. A. Laport; Secretary, W. A. Nichols, Canadian Broadcasting Corp., 1012 Keefer Bldg., Montreal, Que.

NEW ORLEANS—Chairman, G. H. Peirce; Secretary, D. W. Bowman, 8327 Sycamore St., New Orleans, La.

PHILADELPHIA—Chairman, C. C. Chambers; Secretary, R. L. Snyder, 103 Franklin Rd., Glassboro, N. J.

PITTSBURGH—Chairman, P. N. Bossart; Secretary, Ralph Overholt, Jr., 7211 Thomas Blvd., Pittsburgh, Pa.

PORTLAND—Chairman, E. R. Meissner; Secretary, L. M. Belleville, 361 N. E. 75th Ave., Portland, Ore.

ROCHESTER—Chairman, H. J. Klumb; Secretary, O. L. Angevine, Jr., Stromberg-Carlson Telephone Manufacturing Co., Rochester, N. Y.

SAN FRANCISCO—Chairman, L. J. Black; Secretary, Karl Spangenberg, Electrical Engineering Dept., Stanford University, Calif.

SEATTLE—Chairman, K. H. Ellerbeck; Secretary, R. J. Gleason, Pacific Alaska Airways, Inc., 3314 White Bldg., Seattle, Wash.

TORONTO—Chairman, R. H. Klingelhoeffer; Secretary, L. C. Simmonds, 301 King St., E., Toronto, Ont.

TWIN CITIES—Chairman, C. A. Culver; Secretary, N. B. Coil, 1664 Thomas St., St. Paul, Minn.

WASHINGTON—Chairman, M. H. Biser; Secretary, C. M. Hunt, WJSV, 817 Earle Bldg., Washington, D. C.

THE INSTITUTE serves those interested in radio and allied electrical communication fields through the presentation and publication of technical material.

THE PROCEEDINGS has been published without interruption since 1913 when its first issue appeared. The contents of each paper published in the PROCEEDINGS are the responsibility of the author and are not binding on the Institute or its members. Material appearing in it may be reprinted or abstracted in other publications on the express condition that specific reference shall be made to its original appearance in the PROCEEDINGS. Illustrations of any variety, however, may not be reproduced without specific permission from the Institute.

STANDARDS reports are published from time to time and are sent to each member without charge. The four current reports are on Electroacoustics, Electronics, Radio Receivers, and Radio Transmitters and were published in 1938.

MEMBERSHIP has grown from a few dozen in 1912 to more than five thousand. Practically every country in the world in which radio engineers may be found is represented in our roster. Dues for the five grades of membership range from $3.00 to $10.00 per year. The PROCEEDINGS is sent to each member without further payment.

SECTIONS in twenty-six cities in the United States, Canada, and Argentina hold regular meetings. The chairmen and secretaries of these sections are listed on the opposite page.

SUBSCRIPTIONS are accepted for the PROCEEDINGS at $10.00 per year in the United States of America, its possessions, and Canada; when college and public libraries order direct from the Institute, the subscription price is $5.00. For other countries there is additional charge of $1.00.

The Institute of Radio Engineers, Inc.
Harold P. Westman, Secretary
330 West 42nd Street
New York, N.Y.

The Institute of Radio Engineers

Incorporated

330 West 42nd Street, New York, N.Y.

To the Board of Directors

Gentlemen:

I hereby make application for ASSOCIATE membership in the Institute of Radio Engineers on the basis of my training and professional experience given herewith, and refer to the sponsors named below who are personally familiar with my work.

I certify that the statements made in the record of my training and professional experience are correct, and agree if elected, that I shall be governed by the Constitution of the Institute as long as I continue a member. Furthermore I agree to promote the objects of the Institute so far as shall be in my power.

....
(Sign with pen)

....
(Address for mail)

....
(City and State)

....
(Date)

SPONSORS

(Signatures not required here)

Mr. ....

Address ....

City and State ....

Mr. ....

Address ....

City and State ....

Mr. ....

Address ....

City and State ....

REQUIREMENTS for Associate membership are that an applicant must be at least twenty-one years of age, of good character, and be interested in or connected with the study or application of radio science or the radio arts.

SPONSORS who are familiar with the work of the applicant must be named. Preferably these should be three Associates, Members, or Fellows of the Institute. In cases where the applicant is so located as not to be known to the required number of member sponsors, the names of responsible nonmembers may be given.

DUES for Associate membership are six dollars per year. The entrance fee for this grade is three dollars and should accompany the application.

(Typewriting preferred in filling in this form) No. ....

RECORD OF TRAINING AND PROFESSIONAL EXPERIENCE

Name ....
(Give full name, last name first)

Present Occupation ....
(Title and name of concern)

Business Address ....

Home Address ....

Place of Birth .... Date of Birth .... Age ......

Education ....

Degree ....
(College) (Date received)

TRAINING AND PROFESSIONAL EXPERIENCE

(Give dates and type of work, including details of present activities)

Record may be continued on other sheets this size if space is insufficient.

Receipt Acknowledged .... Elected .... Notified ....

OTHER GRADES are available to qualified applicants. Those who are between the ages of eighteen and twenty-one may apply for Junior grade. Student membership is for full-time students in engineering or science courses in colleges granting degrees as a result of a four-year course. A special application blank is provided and requires the signature of a faculty member as the sole sponsor. Member grade is open to older engineers with several years of experience. Fellow grade is by invitation only. Information and application blanks for these grades may be obtained from the Institute.

BACK COPIES of the PROCEEDINGS may be purchased at $1.00 per copy where available. Members of the Institute in good standing are entitled to a twenty-five per cent discount.

VOLUMES, bound in blue buckram, may be purchased for $14.25; $11.25 to members.

BINDERS are $1.50 each. The volume number or the member's name will be stamped in gold for fifty cents additional.

INSTITUTE EMBLEMS of fourteen-carat gold with gold lettering on an enameled background are available. The lapel button is $2.75; the lapel pin with safety catch is $3.00; and the watch charm is $5.00. All of these are mailed postpaid.

The Synthetic Production and Control of Acoustic Phenomena by a Magnetic Recording System*

S. K. WOLF†, ASSOCIATE, I.R.E.

Summary—*In recent years there has been an increasingly active search for an electroacoustic system for producing and controlling reverberation and associated phenomena. This paper describes an electromagnetic method of producing and controlling reverberation by the use of a magnetic tape recording system. It consists of recording a sound pattern magnetically on steel tape. The signal is picked up from the tape at frequent split-second intervals and reproduced at any desired level or characteristic. The tape is arranged for driving in an endless helical loop. An obliterating head which continuously obliterates the record is placed just before the first recording head. The phenomena of reverberations and the various methods which have been suggested by others for controlling reverberation synthetically, such as the electro-optical, electromechanical, mechanical-recording, and the reverberation-chamber methods, are briefly discussed. The paper also outlines other uses for the magnetic tape system in the study of acoustic phenomena both synthetically and analytically.*

SINCE the fundamental work of Sabine,[1] most acoustic engineers have, in accordance with the principles he established, solved their acoustic problems by what may be called the "architectural method." There have been improvements in sound-absorbing materials and a few innovations such as the use of the "live end—dead end" in studios, "balanced absorption," "distributed treatment," and "automatically movable surfaces with variable absorption" but there has been no basically new approach to the problem on a practical scale. The Sabine concept of the proper quantity and quality of absorption offers a more or less satisfactory solution to most architectural acoustic problems, particularly where the enclosed space is to be used for one purpose under a given set of conditions.

However, with the advent of broadcasting and recording studios, more flexible and more exacting acoustic conditions are required, and the "architectural principle" of control can, at best, offer only a compromise solution. In auditorium design, architects and engineers alike are still faced with the dilemma of designing the interior ideal for one purpose or resort to a compromise design if the space is to be used for a variety of purposes.

The complex behavior of sound in enclosures has been treated at great length by acoustic engineers. One of the most significant studies was made by Wente[2] in which he treats a room as analogous to a transmission line. Fig. 1 shows the acoustic transmission characteristics of an architecturally treated and an untreated room. This is ample evidence of the complex nature and behavior of sound in enclosures. Curve B shows the influence of "architectural control" on the transmission characteristics.

It is generally accepted that a certain amount of architectural reinforcement of sound energy is desirable. This is probably due to the harmonic nature of speech and music. The amount and kind of influence

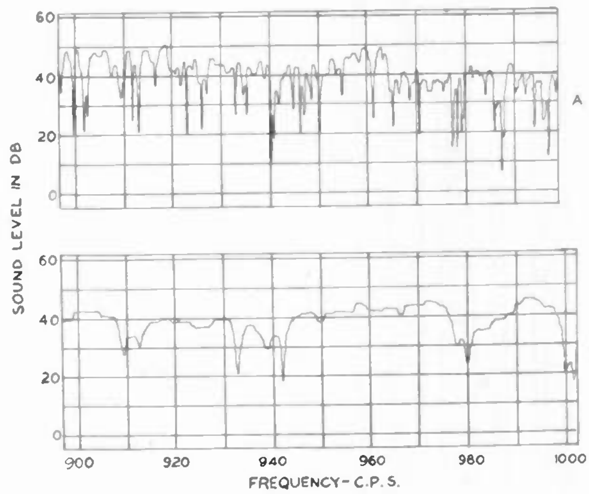


Fig. 1

depends upon the use to which the auditorium is to be put. Even as yet, we have no very accurate way of measuring or prescribing the acoustic excellence of an auditorium. The most universal measure of the acoustic properties of an auditorium is its reverberation time which represents the time required for sound energy to decay 60 decibels, which is rarely, if ever, realized in practice, and furthermore, does not take into consideration the character of the decay. We sometimes measure the number of irregularities in a sound-transmission curve of an auditorium and interpret too many as harmful and a reasonable number as beneficial, but we are unable to separate the desirable from the undesirable.

It is not practicable to control architecturally at will the acoustic characteristics of a space. Such control, however, is desirable in auditoria as well as in recording and broadcasting studios. On the other hand, it is highly probable that we can produce an electrical system whereby we will be able to create synthetically a counterpart of practically any interior. Furthermore, such a device may even yield a space pattern more satisfying than any yet achieved architecturally. In employing the synthetic scheme which I would propose, the auditorium or studio would first be rendered *acoustically inert*, or relatively so, and the sound pattern most desirable would be created electrically.

* Decimal classification: 534×621.385.97. The original manuscript of this invited paper was received by the Institute, June 6, 1941.

† President, Acoustic Consultants, Inc., New York, N. Y., and New York Representative, The Brush Development Company, Magnetic Tape Division, Cleveland, Ohio.

[1] Wallace C. Sabine, "Collected Papers on Acoustics," Harvard University Press, Cambridge, Mass., 1927.

[2] E. C. Wente, "The characteristics of sound transmission in rooms," *Jour. Acous. Soc. Amer.*, vol. 6, p. 121; 1935.

This basically different approach is not new. It has been suggested by several experimenters with as many methods. But there is yet to be developed a completely practical system for solving most acoustic problems by the synthetic method. Presently, a system will be described which, the author believes, is inherently capable of satisfying all the requirements. Before describing this system, some of the more important synthetic methods that have been suggested will be discussed briefly. They may be classified as architectural, mechanical, and electrical. The last may be subdivided into the electric-delay circuit, the electromechanical, the electrooptical, and the electromagnetic systems.

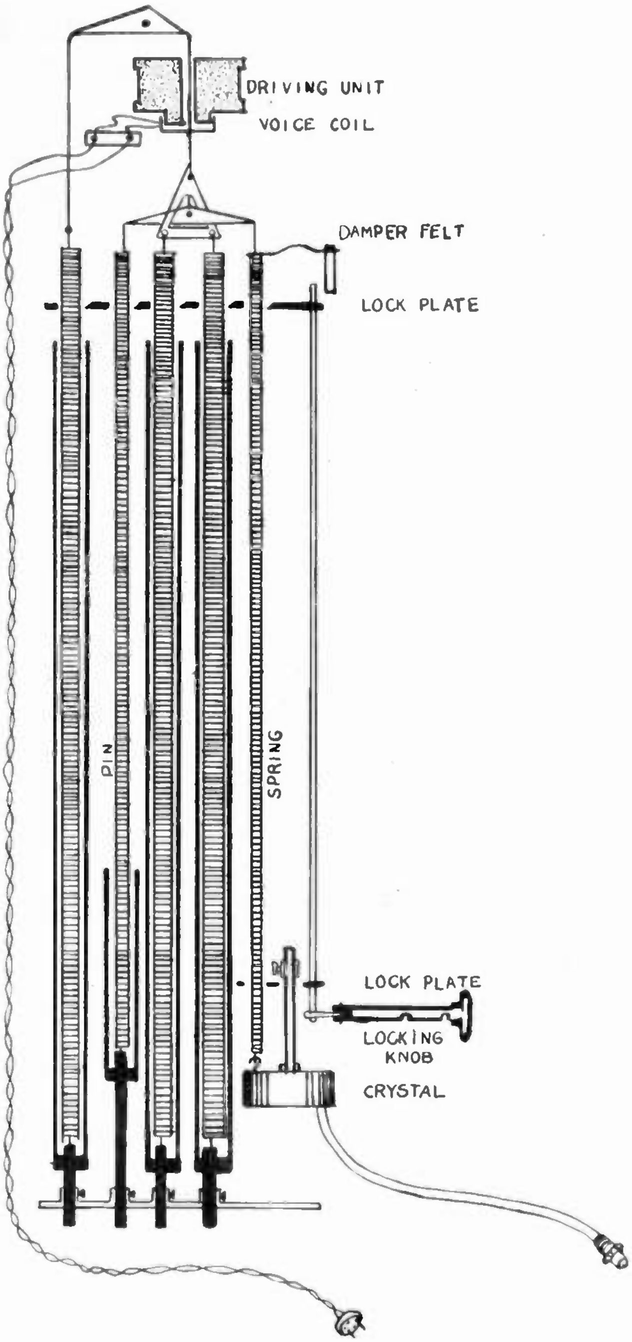


Fig. 2—Reverberation unit.

The methods employ either a capacity storage or storage by recording.

The echo chamber, which has been classified as architectural, is an acoustic capacity-storage device and is perhaps the most widely used method of controlling reverberation in motion-picture recording[3] and radio broadcasting[4] studios. Today echo chambers vary from large empty rooms of 10,000 or more cubic feet to narrow labyrinthlike chambers in which loud speakers and microphones are placed for introducing acoustic energy and microphones for picking up the reverberant energy.

The surfaces of such chambers are covered with glazed, hard-surfaced materials and have reverberation time periods from 2 to 5 seconds. Control is accomplished by sending any desired part of the signal through the echo chambers before recording or broadcasting. The cost and control limitations of echo chambers make them impractical for very extensive use.

Electrical delay circuits for controlling reverberation was suggested by Mills.[5] This method consists in transmitting electrical energy over an artificial line, reflecting it at a distance, and transmitting it back again, thus obtaining twice the phase shift or delay which could be obtained by transmitting them over the line in only one direction.

An electromechanical, or spring system, of reverberation control was developed by Laurens Hammond.[6] The unit consists of a small electromechanical device a few square inches in cross section and approximately 4 feet in length and is shown in Fig. 2.

The energy passing through the unit drives a small moving coil of the same type and general design as is normally used to operate a dynamic loud speaker. When the coil moves, it does not radiate any appreciable sound, but instead it actuates a very small aluminum cage to which are fastened a multiplicity of small helical steel springs and levers. One of these springs in turn actuates a Rochelle-salt crystal. The latter produces a new electrical signal which is amplified. The velocity of sound in the spring is of the order of 50 feet per second. To produce a partial reflection point, the spring is joined to another section of spring of slightly different size resulting in the desired "mismatch." Damping of the spring is accomplished by terminating it in oil. The number of different lengths of paths from the input to the output is enormous because it rises geometrically from the number of reflecting points.

The Hammond unit is now widely used with electric organs and is suitable for use with electronic musical instruments. Its inexpensiveness recommends it where it can be used, but it is not suitable for speech or general reverberation-control purposes.

[3] John K. Hillard, "Reverberation control in motion picture recording," *Electronics*, vol. 11, p. 15; January, 1938.
[4] Howard A. Chinn, "Reverberation control in broadcasting," *Electronics*, vol. 11, pp. 28–29; May, 1938.
[5] John Mills, U. S. Patent No. 1,647,242.
[6] Laurens Hammond, "Reverberation control with the Hammond organ," published by Hammond Organ Company, 1939.

A disk recording method was suggested by Arnold.[7] It consists of a plurality of disk pickup devices so located with respect to each other and with respect to the record surface to secure proper time intervals of the successive co-operation of each pickup device with the same portion of the record surface, each pickup device being associated with an attenuator or amplifier to secure the proper relative loudness.

An electrooptical method for producing reverberation was developed by Goldmark and Hendricks.[8] The basic principle consists in recording a fugitive sound pattern of an original sound signal on the rim of a rotating phosphor-coated disk by means of a modulated light source and a simple optical system. The signal is picked up from the disk at later points through another optical system and photocells. The logarithmic decay of the sound images on the disk as they pass the phototubes gives reverberation effects. The secondary signal is then mixed with the original signal as in other synthetic-reverberation systems. Fig. 3 shows a schematic diagram of the Goldmark system.

The use of a magnetic wire recording system, or the telegraphone, was first suggested by Alfred N. Goldsmith.[9] His patent outlines in detail several control systems. The use of magnetic tape employing the telegraphone principle as suggested was developed in an experimental form as suggested by Begun[10] and the author.[11]

The original design of a magnetic tape system proved to be impractical because of the noise created by the joint in the endless tape when it passed the reproducing heads. This major problem in the magnetic tape system had to be solved before such a device could be made practical. The engineers of The Brush Development Company solved it and completely redesigned the original experimental model, which will be described in detail.

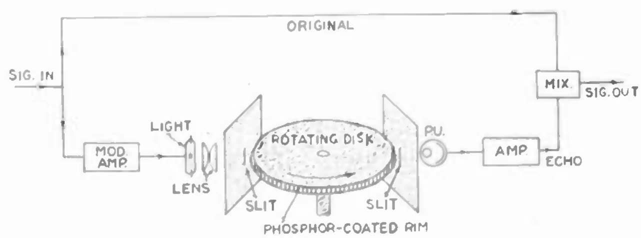


Fig. 3—Basic schematic diagram.

The Brush[12] unit consists essentially of a magnetic tape recording system in which the original sound is amplified, reproduced, and recorded simultaneously. The recorded signal is then reproduced at split-second intervals by a series of pickup heads with diminishing intensity. Each of the pickup devices is delayed in time with respect to the preceding one with reference to the passage of the sound carrier through these devices. These successive reproductions constitute, in effect, a series of echoes of the original sound.

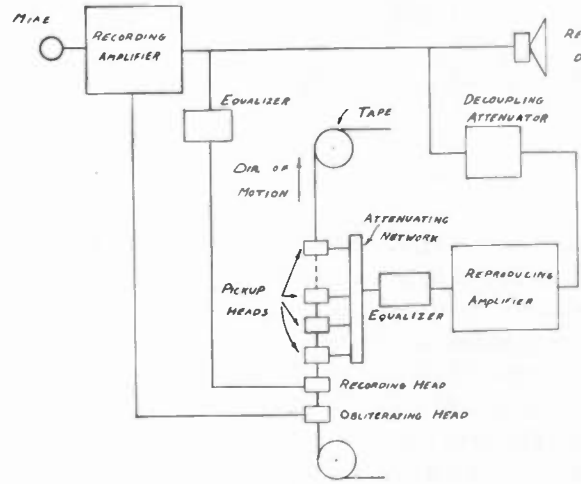


Fig. 4

The block diagram of Fig. 4 shows a schematic diagram of the essential elements of the synthetic-reverberation system. The original sound signal is first passed through an amplifier, the output of which is split into two channels. Through one channel the sound signal passes through an equalizer and goes to the recording head, while through the other, it goes to the output of the system to provide the "direct" sound signal. This amplifier also supplies the obliterating current. After being recorded on the tape, the signal is picked up by a series of reproducing heads. Associated with each head is an attenuating network which is so constructed that the level from successive heads decreases exponentially. The outputs of all of these heads are combined and fed to the input of another amplifier. The output of this playback amplifier is then combined with the direct channel to give the desired reverberation. These are, of course, only the essentials of the machine and give only the one reverberation time for which the attenuating network was calculated. Some of the other factors which are of importance in the actual construction and adjustment of the machine will now be described.

The first consideration is the best spacing of heads to use. The first thought might be that their spacings should not be equal or multiples of any distance; otherwise, the phase relationships of the signals from the various heads might cause all the voltages to add for some frequencies, and to cancel for other frequencies. Practically, it is impossible to obtain absolutely equal spacings, and with the approximately equal spacings realized in practice, this difficulty is not encountered.

Experiments showed that the most important obstacle to be overcome was that the machine tended to

[7] H. D. Arnold, U. S. Patent No. 1,859,423.
[8] Peter C. Goldmark and Paul S. Hendricks, "Synthetic reverberation," Proc. I.R.E., vol. 27, pp. 747–752; December, 1939.
[9] Alfred N. Goldsmith, U. S. Patent No. 2,105,318.
[10] S. J. Begun and S. K. Wolf, "On synthetic reverberation," *Communications*; August, 1938.
[11] S. K. Wolf, "Artificially controlled reverberation," *Jour. Soc. Mot. Pic. Eng.*, pp. 390–397; April, 1939.
[12] E. S. Rich, "Report on a method of obtaining artificial reverberation using a magnetic tape recording system," published by The Brush Development Company, January, 1941.

give a series of distinct echoes rather than a blended reverberation. This effect was most noticeable for speech and for sharp sounds, like clapping of hands or striking metal with a hammer. In an actual room such effects are not noticed because the multiple reflections from the bounding surfaces give a vastly greater number of echoes which are heard at considerably shorter

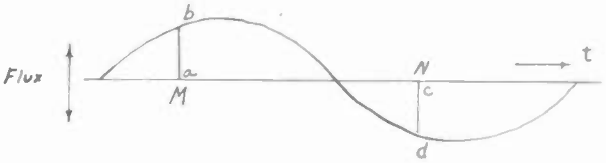


Fig. 5

intervals. Theoretically, then, it would seem that an infinite number of heads would be the obvious solution for producing synthetic reverberation. However, a further analysis shows that this is not the case. A reasonable number of pickup heads spaced along the tape gives the illusion of reverberation rather than a series of echoes. The fact that the ear apparently cannot differentiate between a series of echoes diminishing in intensity and a continuously decaying sound leads one to surmise that the ear may possess an audio persistency similar to the persistency of vision in the eye. Furthermore, it is an interesting fact that the minimum rate at which echoes must occur for "blending" is of the same order of magnitude as the minimum number required for blending of the minimum number of light impulses for visual persistency. Rich[12] observed in his experiments that "a number of pickup heads spaced along the tape gives only one echo at a time, the others following in rapid sequence, and these would correspond to reflections from two parallel walls only, and not from four walls, a ceiling, and a floor. In other words, it is a one dimensional proposition rather than a three dimensional one." However, the illusion is satisfying.

The use of an infinite number of heads spaced infinitely close is undesirable for this application for another reason. If there were no attenuation of one head with respect to another, there would be no output. For every positive voltage developed in one head, there would be an opposite voltage developed in another head by the negative half of the flux wave. Referring to Fig. 5, for a head placed at point M having a voltage developed in it proportional to flux ab, there is a head at point N having a voltage proportional to cd which would cancel it. However, if there were progressive attenuation between successive heads, the voltage from the head at N would not completely cancel that of the head at M. Considering that the attenuation between heads is a function of the displacement, it follows that a low frequency with a long wavelength would suffer less cancellation than a high frequency with a short wavelength. This means that the resultant voltage from all the heads is greater for low frequencies than for high ones and is, in fact, inversely proportional to frequency. Expressed mathematically, this is of the form

$$e = \frac{k}{f}$$

where e is the resultant voltage generated, g, the frequency, and k, a constant. From the similarity of this to the expression for voltage across a condenser

$$e = IX_c = \frac{I}{2\pi fC} = \frac{K'}{C}$$

it is seen that an infinite number of heads with progressive attenuation would correspond to a shunt capacitance across the input of the amplifier. Actually this inherent effect of frequency discrimination is present with a finite number of heads, but is practically negligible.

From experiments, it was found that if the head spacings corresponded to a time delay of 1/20th of a second or less, the individual echoes were reasonably well integrated. In the machine that was built, an interval corresponding to 1/40th of a second was used for the first heads of the series, and a somewhat greater distance was used for the last heads whose outputs were attenuated considerably with respect to those of the first heads.

Again, before being able to determine the number of heads to be used, it was necessary to decide on the reverberation time desired. As mentioned previously, flexibility is one of the desirable qualities for a device of this sort, so it should provide different amounts of reverberation. Obviously a single continuously variable control of reverberation time would be best, but since this means the equivalent of a volume control for each head, all having a different exponential taper and ganged together, it was not considered practical at that stage of the development. A push-button switch giving a choice of 0, $\frac{1}{2}$, $\frac{3}{4}$, 1, or 2 seconds reverberation time seemed the best solution of this problem. These values were determined by experiment, $\frac{1}{2}$ second being quite satisfactory for speech, and 2 seconds being sufficient for a symphony orchestra.

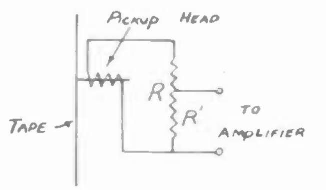


Fig. 6

A long bar with pulleys at each end to guide the tape was provided upon which to mount the heads. The length of the bar and the speed of the tape were such that the maximum time delay obtainable for a series of heads was about $\frac{1}{2}$ second. Knowing this, the attenuation required for the last head is easily determined from the definition of reverberation time. For 2 seconds reverberation time, the level should have dropped 15

decibels in $\frac{1}{2}$ second; for 1 second reverberation time, it should have dropped 30 decibels in $\frac{1}{2}$ second; for $\frac{1}{2}$ second reverberation time, it should have dropped 60 decibels, and so on. It was found that heads which attenuated more than 30 decibels contributed a negligible amount to the over-all quality of the reverberation, and so they were cut out entirely. Knowing the attenuation for the last head for each reverberation time, the attenuation for the other heads is easily determined, since the decay is plotted in decibels linear with time.

Fig. 6 shows the type of attenuation network used, and Fig. 7 shows the taps and switching arrangement used to obtain the different reverberation times. Thirteen playback heads in all were used for this, and their outputs were connected in series to go into the playback amplifier. The attenuation network for each head was built up of 5 small resistors to give the proper taps for the 5 reverberation times.

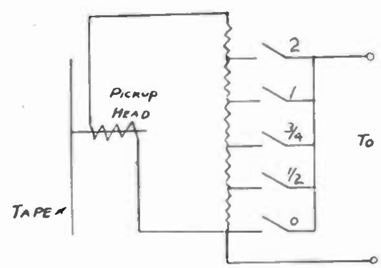


Fig. 7

As was brought out previously in this discussion, producing different decay times for different frequencies would require a different response from each pickup head. To obtain this would require a separate and different equalizer for each head. Furthermore, as was also brought out, it is desirable to have this frequency response variable to give the effect of rooms with different absorption characteristics. The design of a separate variable equalizer for each head was left for future consideration. In the machine built, a compromise was made in this respect, and all heads were adjusted for the same frequency response, and a single variable equalizer was used. This was the schematic drawing of the circuit shown in Fig. 8. With the control in the

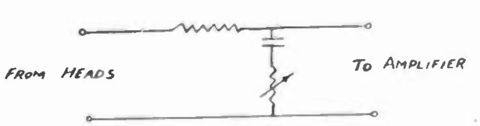


Fig. 8

position for maximum high-frequency level, the response obtained was such that the level dropped gradually for frequencies above about 800 cycles, being 18 or 20 decibels down at 6000 cycles. Causing the low frequencies to have a higher level than the high frequencies gives an effect of accentuated bass notes to the reverberation.

It has been previously mentioned in this article that earlier designs of magnetic tape reverberation synthesizers were impractical because of the fact that there was a series of loud clicks in the output each time the tape splice passed through the heads. These clicks were so objectionable that a method had to be devised to eliminate them. The method used consisted in cutting off the playback amplifier for a fraction of a second while the joint passed through the heads. Since the direct channel was not interrupted, the sound signal was

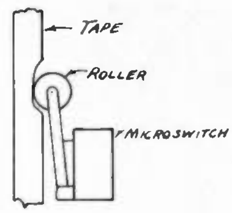


Fig. 9

not lost altogether—only the reverberation. To accomplish this, a microswitch actuated by a lever with a roller at one end was mounted so that the roller was in contact with the edge of the tape. This arrangement was designed by the engineers of The Brush Development Company. A notch about $\frac{3}{4}$ of an inch long was

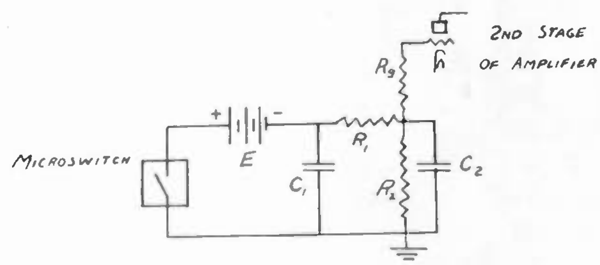


Fig. 10

cut in the tape as shown in Fig. 9, so that the switch would be actuated just before the tape joint reached the first pickup head. The operation of this switch applied a bias to the second stage of the amplifier sufficient to make the tube inoperative. The circuit of Fig. 10 was used so that this sudden application of voltage would not in itself cause an objectionable click.

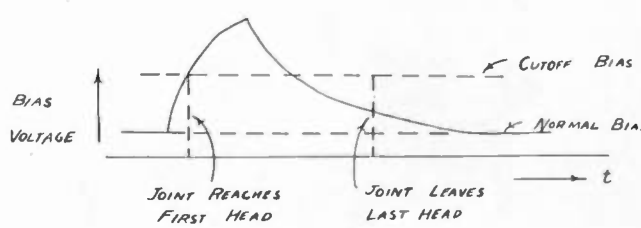


Fig. 11

When the microswitch is closed, condenser C_1 is immediately charged to the full voltage of the direct-current supply E. This charges condenser C_2 through resistance R_1, and thus puts additional bias on the tube. Condenser C_2 is then discharged through resistance R_2 and the amplifier is again in operation. The time constants of the two circuits are chosen so that there is a rapid build-up of biasing voltage, and a slower decay. The relation of this to the passage of the joint through the heads is shown in Fig. 11. As can be seen, the am-

plifier is already slightly operative before the joint leaves the last heads, but this is not objectionable because the output of the heads is attenuated considerably with respect to the direct level. Since the amplifier is blocked for less than $\frac{3}{8}$ of a second, the temporary loss of the reverberation is not noticeable.

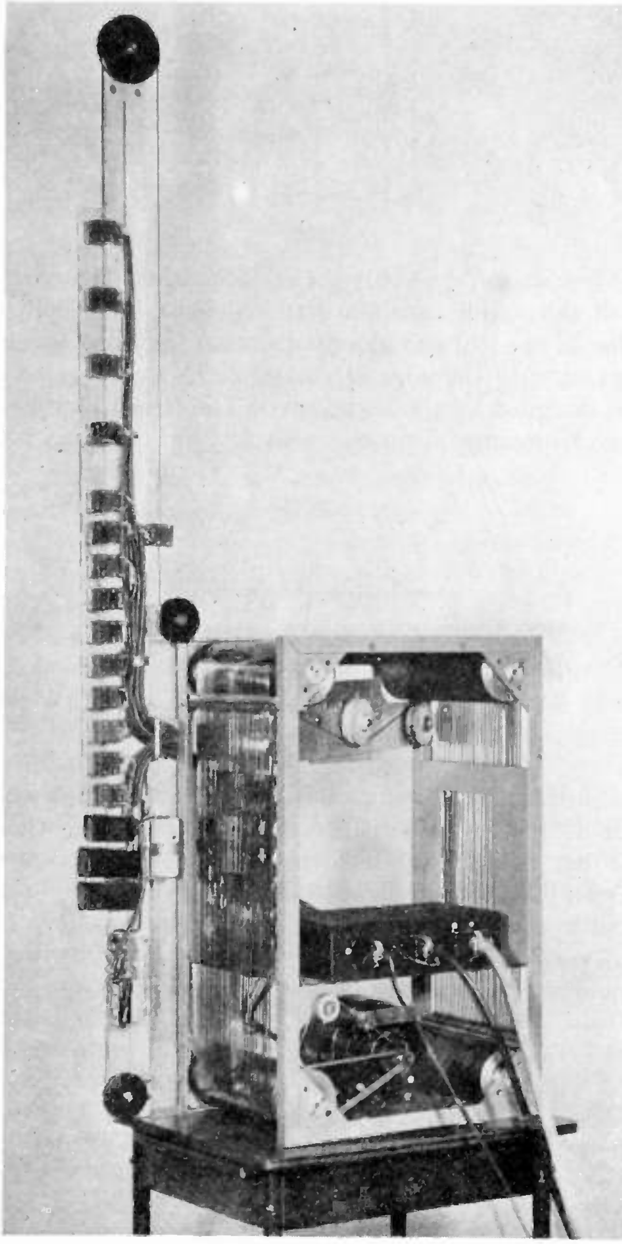


Fig. 12

The method of coupling the outputs of the two amplifiers should also be mentioned. Since this machine was designed to operate with regular studio amplifiers, no special attempt was made to obtain sufficient power output to drive a speaker. As the recording head was connected to the output stage of the recording amplifier, it was necessary, in order to prevent feedback through the tape, to place an attenuator pad between the output of the playback amplifier and that of the recording amplifier when mixing their outputs. By means of a semifixed volume control on the chassis of the playback amplifier, the level of the signal from this amplifier at the output terminals was adjusted to its proper value with respect to the direct level at the same terminals, and thereafter was not changed, the studio amplifier being used to give all control of the over-all volume.

Fig. 12 is a photograph of the tape unit showing the spacing of the heads. Fig. 13 shows the control unit.

The illusion of reverberation produced by this unit is very realistic. Both speech and music attain more pleasing qualities when the proper amount of reverberation is added. Individual echoes are almost unnoticeable, but instead, they are blended into a gradual decay of sound. The greatest value of the unit seems to lie in the great increase in brilliance and color which is given to music. This is especially true for music of large orchestras, the music of which had been picked up by microphones giving too little reverberation.

A machine like the one described might easily be modified so that other sound effects could be obtained as well. For instance, a single pickup head or a series of widely spaced heads would give either a single echo, or a series of distinct echoes. Unusual effects would be produced by varying the level of the echo, making it, for example, louder than the direct sound.

If a synthetic-reverberation system is to become generally used in auditoria and recording and broadcasting studios, it must be useful for speech, music, and sound

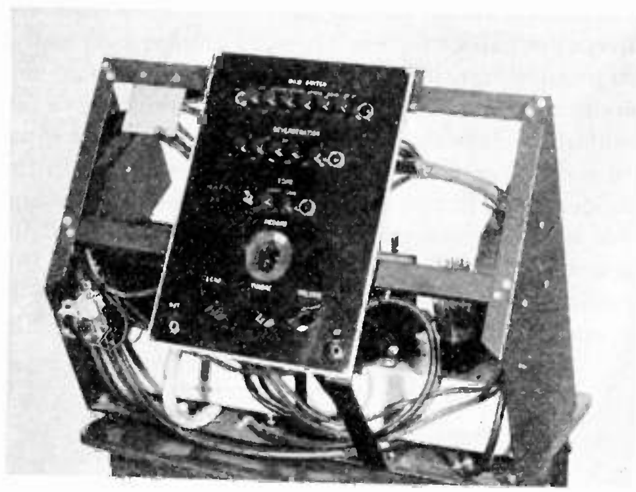


Fig. 13

effects, must be mechanically and electrically foolproof, have low maintenance cost, and must make it possible to control the intensity and frequency response of the signal.

At the present time, about the only system that gives promise of meeting all of these requirements is a magnetic tape recording system. This device has been used experimentally by the three large broadcasting systems and in a theatrical production at Stevens

Institute in Hoboken, New Jersey, the results of which were discussed by Burris-Meyer.[13]

In addition to reverberation and echo control, the magnetic tape system should prove to be a valuable experimental tool in the field of sound research. The following are some of the more obvious experimental uses: (1) the study of acoustical patterns; (2) the study of the modulation of sound-decay curves; (3) the study of desired frequency response; (4) the study of the build-up and decay of acoustic energy in auditoria; and (5) the study of space phasing for improving the naturalness of the public-address system.

Vincent Mallory, Harold Burris-Meyer, and the author conducted some preliminary experiments on the last-mentioned application and can report that a higher degree of intimacy and intelligibility can be accomplished by such a system than is accomplished by present-day public-address systems.

[13] Harold Burris-Meyer, "The use of the remade voice subsonic and reverberation control." Presented, Acoustical Society of America, Rochester, N. Y., May, 1941.

Television Transmission*

M. E. STRIEBY†, MEMBER, I.R.E., AND C. L. WEIS†, ASSOCIATE, I.R.E.

Summary—*Experiments in the transmission of television signals over wire lines have been made from time to time as the television art has developed.[1-9] The present paper discusses experiments made during the summer of 1940 with 441-line, 30-frame interlaced signals transmitted over coaxial cable and other telephone facilities. Some of the general problems of wire transmission have been included. In particular, the results of transmission studies on a system linking New York and Philadelphia will be reported.*

Source of Signal and Receiving Device

As a source of television signals for testing purposes it was desirable to obtain high-grade signals covering a variety of images. Among the things desired were a high signal-to-noise ratio, faithful reproduction of a wide range of tones from black to white, a high degree of freedom from distortions in response over the frequency band, and an adequate frequency band. A moving-picture projector with an image-dissector tube and the necessary circuits for providing sweep, blanking, and synchronization were assembled for this purpose and are described by Jensen.[7] This picture-signal generator is located in the Bell Laboratories at 180 Varick Street. Signals generated there were transmitted to the Pennsylvanian Hotel over telephone cable circuits equipped with suitable amplifiers and equalizers, using methods discussed later. The results show that a signal has been obtained with reasonable freedom from noise and distortion, of high contrast, and a band width of about 4 megacycles. A complete carrier system over the coaxial cable to Philadelphia and back was added to the video circuit. Using the repeaters which were designed primarily for 480-channel telephone service this system provides an effective band with of about $2\frac{3}{4}$ megacycles.

For reproduction of transmitted scenes, a special tube[10,11] was provided in order that those viewing the tests might be able to see troubles or distortion in the system more clearly than would be seen on present-day commercial receiving sets. This was accomplished by a number of refinements which are justified by the desire for a high-grade testing tool. A very long tube (about 5 feet) was built to avoid moving the electron beam over a wide angle. Ten-thousand-volt beam potential and an efficient screen material were used to obtain a high-light brightness of 10 foot-lamberts with a picture area of 0.5 square foot. Accurate overlap and interlacing were obtained by using a rectangular spot and carefully designed sweep circuits. The area, rather than the intensity, of this spot is modulated by the signal to control the picture brightness. At maximum brightness the electron beam, which is square in cross section, passes through a rectangular aperture of about the same size. At minimum brightness the beam is moved horizontally so that none of it, except stray electrons, passes through the aperture. This results in a rectangular spot of constant height and a width determined by the amount of the modulating signal. This design gives a tube which is characterized by the absence of blooming, and by an improvement in the sharpness of focus at the edges of the field.

In order to reproduce a good picture it is necessary to reproduce small incremental changes in brightness

* Decimal classification: R583. Original manuscript received by the Institute, April 8, 1941. Presented, Sixteenth Annual Convention, New York, N. Y., January 11, 1941.
† Bell Telephone Laboratories, Inc., New York, N. Y.
[1] Herbert E. Ives, "Television," *Bell Sys. Tech. Jour.*, vol. 6, pp. 551–559, October, 1927.
[2] Herbert E. Ives, "Image transmission system for two-way television." *Bell Sys. Tech. Jour.*, vol. 9, pp. 448–469; July, 1930.
[3] R. D. Kell, A. V. Bedford, and M. A. Trainer, "An experimental television system—the transmitter," Proc. I.R.E., vol. 22, pp. 1246–1265; November, 1934.
[4] "The first real television. How the coronation procession was televised," *London Telev.*, vol. 10, pp. 335–339; June, 1937.
[5] "Broadcasting and the coronation," *Nature*, vol. 139, pp. 747; May, 1937.
[6] "Der Deutsch-Sprechverkehr eroffnet," *Berlin Leipzig Telev.*, May 25, 1936.
[7] Axel G. Jensen, "Film scanner for use in television transmission tests," Proc. I.R.E., vol. 29, pp. 243–250, May, 1941.
[8] "Television-Telephone," "Berlin-Munich Television," *London Telev.*, vol. 11, p. 468; August, 1938.
[9] *Revue des telephone, Telegraphes et T.S.F.*—"L'overture due Service de Visiotelephonie," vol. 16, p. 343, April, 1939.
[10] C. J. Davisson, United States Patents, Nos. 2,217,197 and 2,217,198, October 8, 1940.
[11] H. E. Ives, "Transmission of motion pictures over a coaxial cable," *Jour. Soc. Mot. Pic. Eng.*, vol., 31 pp. 256–272; September, 1938.

as well as to provide a wide range of picture brightness. By adding a compensating circuit at the signal generator which corrects for the nonlinearity of the receiving

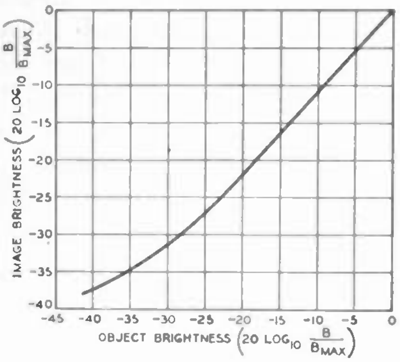


Fig. 1—Brightness range of system.

tube, it has been possible to increase, by about 3 times, the range of object brightnesses for which significant small steps are discernible. The over-all object—image brightness characteristic for the signal generator and the receiving tube is shown on Fig. 1. This compensation materially improves the picture without imposing more rigid requirements on the transmitting medium or the receiving device.

Coaxial Cable Systems

The general problem of telephone and television transmission over long coaxial cable systems has been discussed previously[12,13] and will be dealt with here only to the extent necessary to describe the presently available facilities. The type of cable now being engineered[14] represents some improvements over the New York-Philadelphia cable used for the demonstration and previously described. The new type of cable[15] is illustrated in Fig. 2, which shows a section of the

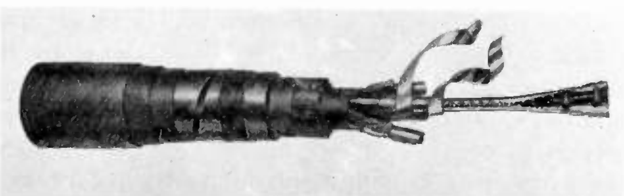


Fig. 2—Photograph of coaxial cable—Stevens Point-Minneapolis type.

underground cable which was installed last year between Stevens Point, Wisconsin, and Minneapolis, Minnesota, a distance of about 200 miles. It contains four coaxial conductors, each about $\frac{1}{4}$ inch in diameter. Two of these are used for regular transmission, one in each direction. The other two are completely equipped and carry the same signal in a parallel emergency circuit. Arrangements are provided to cut in the emergency circuits automatically at intervals of 50 miles in case of trouble on the regular circuits. The measured transmission characteristics of these conductors are shown in Fig. 3.

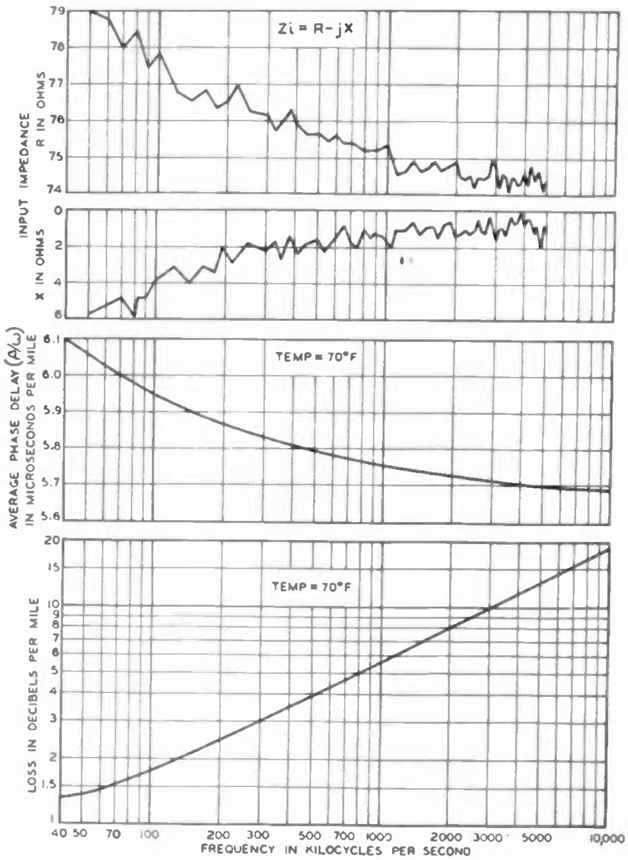


Fig. 3—Constants of 0.267 inch coaxial cable.

Repeaters

The band width which a coaxial system will pass is determined by the design of the amplifiers which are provided to overcome the attentuation. The system used for these tests was equipped with "3-megacycle" amplifiers which are installed at intervals of about 5 miles—a total of 40 in the demonstration loop circuit from New York to Philadelphia and return. As can be seen in Fig. 3, the attenuation of 5 miles of present-design coaxial cable is about 50 decibels at 3 megacycles. This loss is quite accurately counterbalanced by the gain of the amplifier and associated equalizing equipment over a range from about 60 to 3100 kilocycles. The amplifiers themselves are small plug-in units, two of which, one for each direction of transmission, constitute a repeater, and are mounted with the necessary associated equipment in a waterproof housing about 2 feet×2 feet×1 foot. Such a repeater mounted on a pole is shown in Fig. 4. At other loca-

[12] L. Espenschied and M. E. Strieby, "Wide band transmission over coaxial lines," *Elec. Eng.*, vol. 53, pp. 1371–1380; October, 1934.
[13] M. E. Strieby, "A million cycle telephone system," *Elec. Eng.*, vol. 56, pp. 4–7; January, 1937.
[14] O. S. Markuson, "Stevens Point-Minneapolis coaxial cable," *Bell Lab. Rec.*, vol. 19, pp. 138–142; January, 1941.
[15] J. F. Wentz, "Transmission characteristics of the coaxial structure," *Bell Lab. Rec.*, vol. 26, pp. 196–200; February, 1938.

tions these are placed underground in manholes, or in suitable shelters.

Due to temperature variation the attenuation of the cable changes by about ± 10 per cent during the year, if it is installed above ground, or about one third that amount if it is placed underground in the usual manner. These changes are compensated over the whole frequency range by an automatic regulating device associated with each amplifier and actuated by a pilot channel at 2064 kilocycles. At the output of each amplifier this pilot frequency is selected, amplified, and used in a back-acting circuit to control the amplifier gain properly over the frequency range. This device maintains the transmission even on long systems to within a fraction of a decibel near the pilot frequency and to a fair degree of approximation over the entire frequency range. As the length of the system is built up it has been found desirable to provide mop-up equalization at intervals of about 50 miles to neutralize the cumulative effect of these inaccuracies. As a result it is feasible to keep the transmission over the entire frequency band within a fraction of a decibel for a number of such circuit lengths. For circuits of 500

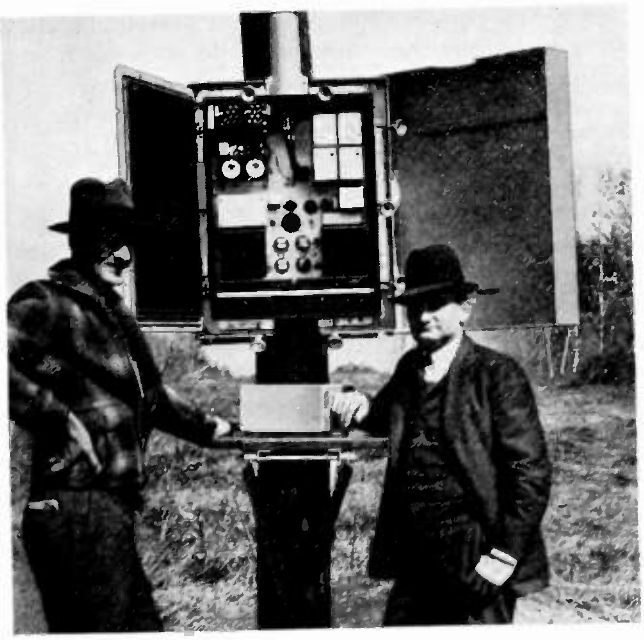


Fig. 4—Photograph of coaxial repeater.

miles or more in length it is expected that it will be necessary to do further mopping-up and adjusting from time to time to take care of the accumulation of these distortions. As an aid in maintaining the system, two additional pilot channels have been provided, one at 64 kilocycles and another at 3096 kilocycles. At each 50-mile attended station these actuate voltmeters to indicate the transmission at these frequencies.

Power for operating the amplifiers is fed over the coaxial conductors themselves at 60 cycles from attended stations about 50 miles apart. Each repeater, including the automatic regulators, uses about 50 watts. The commercial power supply at attended stations is supplemented by stand-by equipment for emergency use in case of commercial power failure.

Noise and Modulation

Extraneous noise from the line, such as static, radio interference, and cross talk, is practically negligible above about 50 kilocycles in this design of coaxial

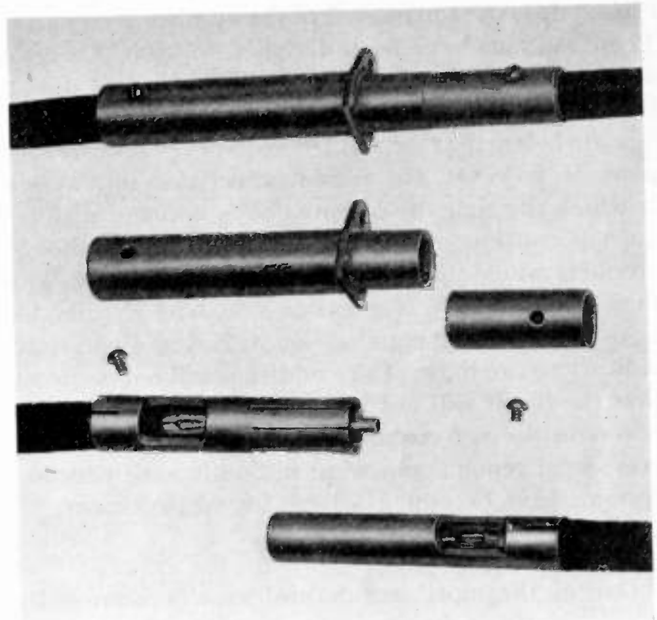


Fig. 5—Photograph of coaxial plug and jack.

cable. Effective shielding of the repeaters and terminal apparatus is necessary to prevent outside interference from entering the system at these points. As far as possible all parts of the circuit carrying the signal are connected by coaxial leads until inside well-shielded compartments. Thus we may visualize the system as one continuous coaxial line with bulges introduced wherever equipment is added. Patching connections are made by coaxial plugs and jacks and flexible coaxial cords having heavy braided shields. Fig. 5 shows their construction. Residual noise in the over-all system arises largely from tube noise in the amplifiers and is 2 or 3 decibels above thermal noise over the useful frequency band. Experiments show that this type of noise increases in proportion to the square root of the length of the system.

No appreciable modulation has been found in the cable itself. Modulation is always present in amplifiers, however, and if large enough gives rise to several defects in the television image. As the operating power in this system comes from an alternating-current source, small amounts of 60- and 120-cycle sidebands are present in the received signal. Unless kept at a low level these will appear on the screen, as one or two dark horizontal bands across the picture. The various power sources encountered in long circuits will, in general, not be synchronized with one another or with

the sweep circuits of the receiver; hence the bands will tend to move slowly up and down across the picture. Under normal conditions no effect of this nature can be seen on the 200-mile demonstration circuit.

Another type of modulation trouble may arise from interaction of the various signal components with the pilot channels, the carrier frequency, or between each other. Second-order products have been measured on these systems and are found to increase, as would be expected, as the square root of the number of repeaters. These systems have been designed to keep this effect at a satisfactorily low value. Disturbances due to third-order modulation products on the 200-mile system are below the threshold of visibility. On theoretical grounds, however, one would expect that in a system in which the time of transmission is uniform with frequency continuously along the line the third-order products would add as voltages. If this were the case they would be objectionable on very long circuits, but since we lump the equalization at discrete points the indications are that voltage addition will not occur and that the result will not be so large as to cause serious effects in present coaxial systems. This matter, however, must remain somewhat in doubt until very long systems have become available for measurement.

Delay Distortion

One of the most serious problems encountered in television transmission is to provide equal time of transmission for all of the different frequency components of the signal. The degree to which the transmission time must be equalized is set by the allowable displacement, on the screen, of units of the picture as determined by the ability of the eye to resolve them. The time of transmission of a television system is constant only if the phase shift is proportional to the frequency from the lowest to the highest frequency in the transmitted range. Permissible departures of the phase shift from linearity with the frequency depend principally on whether this departure is abrupt, gradual, or periodic.

On a long repeatered transmission system it is not feasible to measure phase shift; first, because there is no standard of phase reference at the far end and second because it is continually changing with time for all frequencies by small amounts. For this reason it has been found expedient to measure the rate of change of phase shift with frequency over a small interval. This is called an envelope-delay measurement[16,17] and is discussed more fully in a later section.

Requirements on delay for telephone transmission are very lenient and the steps taken to provide a transmission system, good in this respect, are chargeable to television. Differences in delay exist in the cable itself and in most all of the equipment associated with it. Lack of perfect equalization of these delays produces multiple images both preceding and following. Large numbers of such images usually appear as smearing or streamers. Frequency-band limitation also produces transients which are similar to those produced by imperfect equalization of delay and attenuation.

In any system, in which the impedance is not strictly uniform throughout, similar effects are produced by reflections at the impedance irregularities. In the coaxial system, for example, the repeater impedances do not match the cable impedances at all frequencies. The cable itself is not exactly of uniform impedance, due to variations in the manufacturing process. Such impedance irregularities[18–23] give rise to multiple reflections between themselves and result in small portions of the main signal arriving at the far end somewhat delayed, since they have traversed part of the system three times rather than once. These reflected components produce a "ghost" image displaced from the original on the received picture.

On the 200-mile New York-Philadelphia system the effects of the various types of distortion mentioned above are barely visible. In the demonstration their effect on the picture will be exaggerated by removing some of the equalization. With our present manufacturing technique these difficulties do not appear to be serious in the present state of the television art.

Frequency Band Width

It is known that the detail available in a received television image which is free from other defects is determined by the frequency band. The extent to which detail can be appreciated, however, depends upon the eye and other factors. A recent paper,[24] dealing with these matters, shows that for band widths considerably less than 4 megacycles there is a surprisingly small loss of the detail perceptible to the usual observer. This conclusion is illustrated and confirmed by comparison of the image received over the coaxial system between New York and Philadelphia ($2\frac{3}{4}$ megacycles wide) and local transmission (4 megacycles wide). Under usual viewing conditions (4 to 6 feet) many careful observers fail to pick the difference in this comparison. This point is mentioned in order to help appraise the value of band width in relation to its over-all cost.

What band width the economics of television broad-

[16] H. Nyquist and S. Brand, "Measurement of phase distortion," *Bell Sys. Tech. Jour.*, vol. 9, pp. 522–549; July, 1930.

[17] C. E. Lane, "Phase distortion in telephone apparatus," *Bell Sys. Tech. Jour.*, vol. 9, pp. 493–521; July, 1930.

[18] Didlaukis and Kaden, "Impedance uniformity," *Elek. Nach. Tech.*, vol. 14, pp. 13–23; January, 1937.

[19] Pierre Mertz and K. W. Pfleger, "Irregularities in broad-band transmission circuits," *Bell Sys. Tech. Jour.*, vol. 16, pp. 541–559; October, 1937.

[20] L. Brillouin, *Annales des Postes, Telegraphes, et Telephones*, vol. 27, pp. 269–321; April, 1938.

[21] Leon Brillouin, "Irregularities in telephone and television coaxial cables," *Elec. Comm.*, vol. 17, pp. 164–187; October, 1938.

[22] *Annales des Postes, Telegraphes, et Telephones*, vol. 27, pp. 997–1021; December, 1938.

[23] *Annales des Postes, Telegraphes, et Telephones*, vol. 28, pp. 143–173; March, 1939.

[24] M. W. Baldwin, Jr., "The subjective sharpness of simulated television images," *Bell Sys. Tech. Jour.*, vol. 19, pp. 563–586; October, 1940; Proc. I.R.E., vol. 28, pp. 458–468; October, 1940.

casting will justify is beyond our knowledge and the scope of this paper. However, favorable results of the broadcast of the Republican convention, the demonstration accompanying this paper, and the paper by Baldwin,[24] are believed to be contributions to the solution of that problem. Development work to canvass the possibilities of wider-band repeaters and line systems is also under way.

filter for transmission over the coaxial line. The main sideband now extends from 311 kilocycles to approximately 3.3 megacycles with the vestigial sideband extending down to approximately 200 kilocycles. At this point, only partial shaping of the vestigial sideband and carrier-frequency region has been obtained, the remainder of the shaping is accomplished at the receiving terminal.

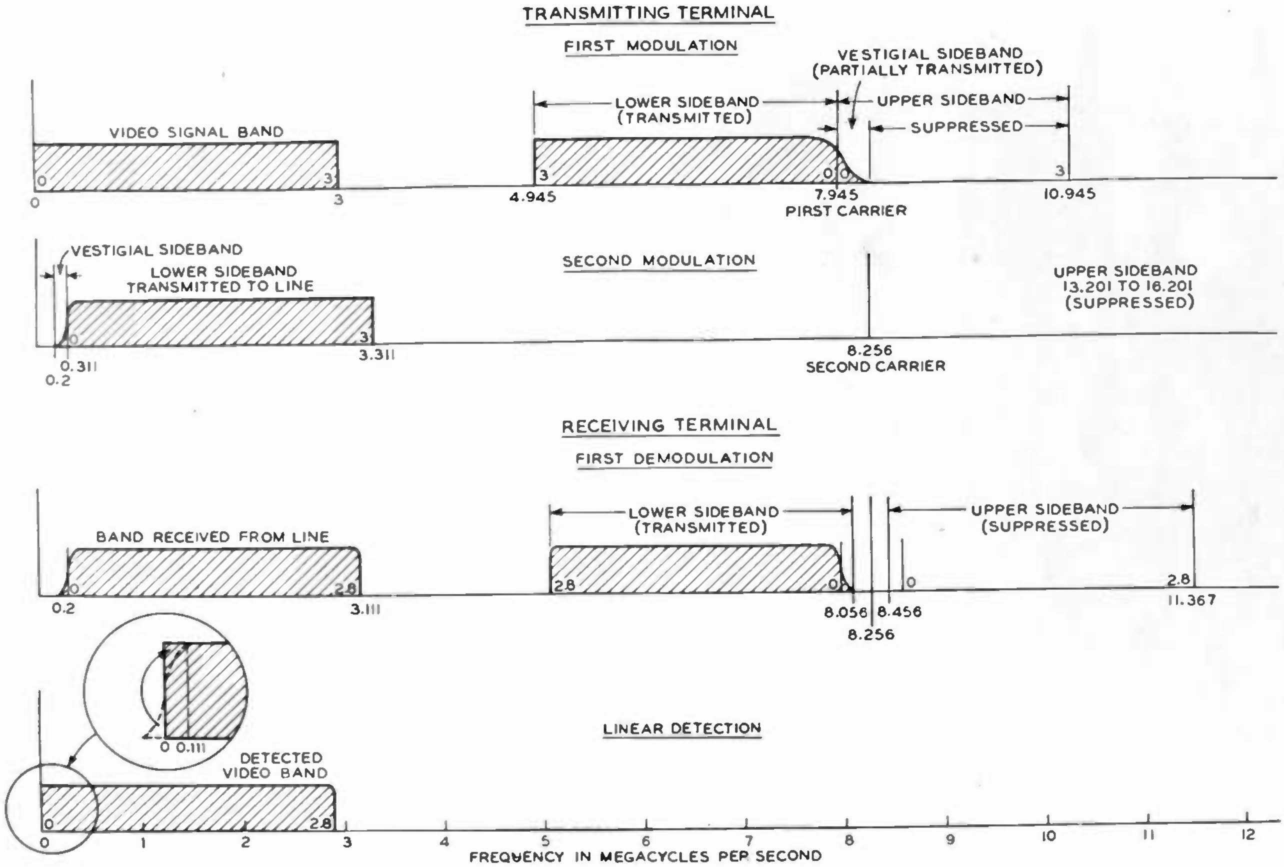


Fig. 6—Method of shifting video to carrier band.

Terminals

For transmission over the coaxial line the video signal must be translated in frequency into the transmission band of the carrier amplifiers. The general method of accomplishing this has been discussed previously.[25] Fig. 6 illustrates the translation method in detail.

Transmitting End

A video signal which extends to 3 megacycles is first modulated with a carrier frequency of 7.945 megacycles. This modulator is of the conjugate-balance type with provisions for transmission of any required amount of average carrier. The lower sideband of this modulation process together with a vestige of the upper sideband is then selected by a band filter. This band is then modulated with a carrier frequency of 8.256 megacycles and the lower sideband of the second modulation process selected by means of a low-pass

[25] M. E. Strieby, "Coaxial cable system for television transmission," *Bell Sys. Tech. Jour.*, vol. 17, pp. 438–457; July, 1938.

There is a concentration of energy in the low-frequency region of a video signal which in this case is in the region of the carrier. It is evident that a considerable advantage, in minimizing modulation distortion and noise introduced by the coaxial line, can be obtained by partially equalizing this energy content over a wider portion of the frequency band. This is accomplished at the transmitting terminal by the use of a "pre-emphasizing" network in which the loss decreases with rising frequency. The delay distortion accompanying the partial shaping of the signal in the transmitting band filter is equalized before transmission over the coaxial line.

The coaxial line utilizes three pilot frequencies for regulation purposes as noted above. The lower pilot frequency, 64 kilocycles, is removed at the receiving terminal by the use of a high-pass filter. The intermediate and upper pilots, 2064 and 3096 kilocycles, respectively, fall within the transmitted television band. The specific frequency allocation was chosen so

as to place these pilot frequencies approximately midway between the line-scanning frequency components. From scanning theory[26] it is known that susceptibility to interference falling midway between scanning-line components is much less than when the interference falls close to them. The residual interfering effect is minimized at the receiving terminal by the use of sharply selective circuits with very little distortion of the adjacent signal components.

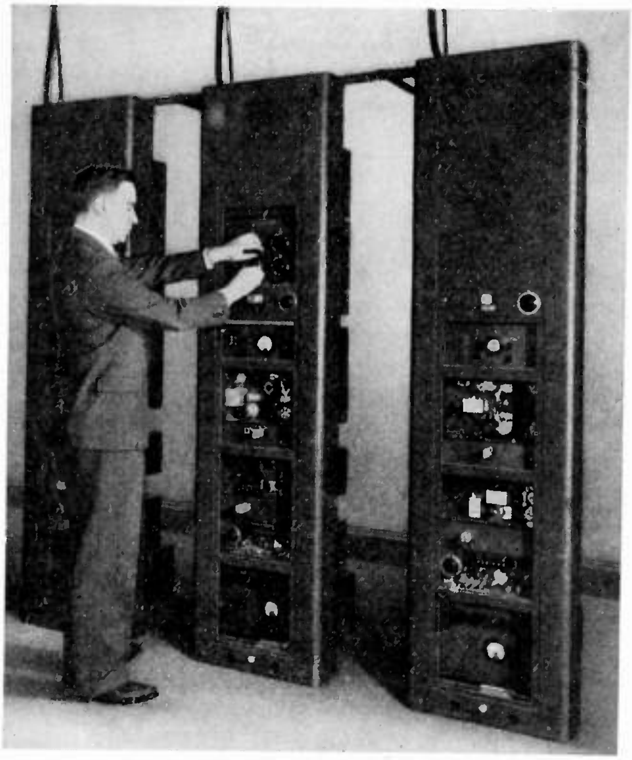


Fig. 7—Photograph of carrier terminal.

Another factor in the choice of line-frequency allocation was the placing of the line-frequency carrier so that the second-order-modulation products also fall in the "empty energy regions" of the television signal. Thus the modulation requirement for second-order products, which is usually the most severe, is reduced considerably.

Provision is made for the transmission of a program channel in the region of 70 to 85 kilocycles; this band is added to and separated from the picture signal at the transmitting and receiving terminals, respectively, by means of appropriate filters.

Receiving End

The function of the television-carrier receiving terminal is to recover the original video signal from the received carrier wave. This terminal is, generally speaking, the inverse of the transmitting equipment. The line signal as received is "restored" or "de-emphasized" by a network having the complementary loss-frequency characteristic of that employed at the transmitter. The signal is then modulated with a carrier frequency of 8.256 megacycles and the lower sideband and a vestige of the upper sideband selected by a following band filter. This filter is a duplicate of the one employed after the first modulation process at the transmitter. At this point the total shaping of the vestigial and main sideband energy in the carrier region has been accomplished so that the addition of the components in these two regions during the final demodulation process will result in the original video signal.

The final demodulation step employs envelope detection using a linear diode rectifier. This method of final demodulation eliminates the necessity of synchronized carrier frequencies at the transmitting and receiving terminals, which we employed in a similar system demonstrated in 1937.[25] The carrier frequencies are supplied from local oscillators, of simple form, which are required to have only moderately stable frequency characteristics.

The "quadrature-component" distortion which accompanies single-sideband or vestigial-sideband transmission systems has been discussed previously.[27] With a ratio of vestigial band width to main sideband width of approximately 4 per cent, the value employed in the present system, this quadrature-distortion term becomes quite large. By transmitting a larger carrier component this distortion term may be reduced to a satisfactory value. In this system the peak-to-peak value of the average transmitted carrier component is made equal to twice the peak-to-peak amplitude of the video envelope of the modulated wave which reduces the distortion from this source to an acceptable amount.

The physical arrangement of the carrier terminals is shown in Fig. 7, where the bay on the right is the transmitting equipment and the left bay contains a complete receiving terminal. Both terminals operate directly from the alternating-current mains.

Video Transmission on Connecting Circuits

The problem of providing short connecting circuits in cities between pickup or distribution points and television studios, control boards, or transmitters is quite different from that of intercity networks. The needs for such circuits may be expected to be extremely variable and to call for circuits on short notice to almost any point. Such facilities may be more or less permanent or may be used only once. These considerations have led to the development of means for adapting an existing telephone plant for television transmission on relatively short notice. For use where permanent facilities can be justified, special types of cable are being developed.

[26] P. Mertz and F. Gray, "A theory of scanning and its relation to the characteristics of the transmitted signal in telephotography and television," *Bell Sys. Tech. Jour.*, vol. 13, pp. 463–515; July, 1934.

[27] H. Nyquist and K. W. Pfleger, "Effect of the quadrature component in single sideband transmission," *Bell Sys. Tech. Jour.*, vol. 19, pp. 63–73; January, 1940.

Television equipment to which such circuits must connect is usually designed to be 72 ohms, unbalanced (i.e., one side grounded) and to use video-signal transmission. To avoid special terminal modulators the signal must be transmitted video. This means transmission of a frequency range of perhaps 45 cycles to 3 or 4 megacycles. For such low frequencies the unbalanced coaxial structure is not well suited, as it is subject to low-frequency interference. For distances of a mile or so a coaxial cable has been used successfully but usually requires special precautions to avoid power-frequency interference. A balanced circuit is to be preferred for this reason.

Telephone Cables

Ordinary telephone cable circuits have been found to be suitable for this purpose. Two difficulties are usually present, first, the high attenuation per unit length and second, the way such cables are branched at frequent points to provide flexibility for regular telephone use. These branches, called "bridged taps," cause such serious impedance irregularities that they must be removed.

The attenuation of ordinary cables ranges from 25 to 95 decibels per mile at 4 megacycles, depending on the gauge and capacitance. Amplifiers are, therefore, required at frequent intervals. The maximum television signal which can be put on such a cable without causing too much interference in adjacent telephone circuits appears to be of the order of 5 volts peak to peak. The lowest level to which the television signal may drop without encountering interference from the telephone circuits varies widely. In areas we have

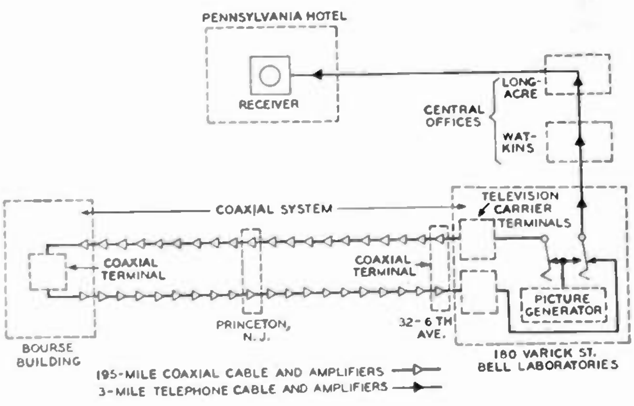


Fig. 8—Map of Pennsylvania Hotel-Varick Street-Philadelphia Circuit.

tested where the circuits are all in cable the lower limit is the order of 65 decibels below 1 volt. However, in a few tests where exposed drop wire enters the cable, the interference was found to be very much higher. The distance between amplifiers in these experiments has varied from about $\frac{1}{2}$ mile to $1\frac{1}{2}$ miles with the average slightly under 1 mile.

The circuit used to bring a demonstration from the Bell Laboratories to the Pennsylvania Hotel used the existing telephone plant and is typical of these facilities. The arrangement is shown in Fig. 8 to consist of three links employing four amplifiers.

Video Amplifiers

Terminal amplifiers are required to provide a means for connecting the 72-ohm, unbalanced, apparatus to the telephone cable which is about 110 ohms, balanced. The receiving amplifier may be seen with the demon-

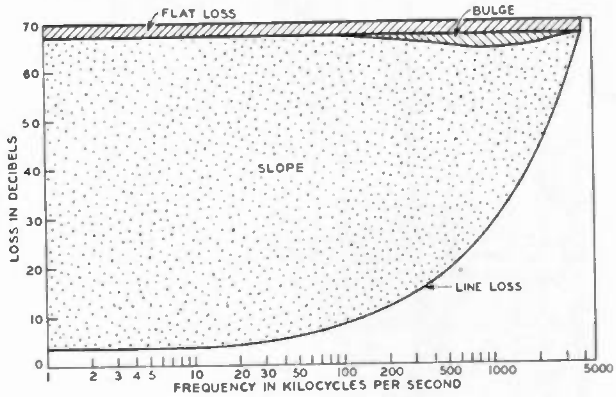


Fig. 9—Equalization method on video circuits.

stration apparatus. These amplifiers are designed to introduce some pre-emphasis of the high frequencies at the sending end and corresponding de-emphasis at the receiving end. This arrangement increases the energy at high frequencies on the line where it is subject to interference from relay operation, particularly in machine-switching telephone offices. The intermediate amplifiers have essentially flat gain—frequency characteristics before the equalizers are added.

The equalization of such circuits for amplitude and delay distortion is accomplished by equalizers at the intermediate and receiving points. The equalizers are variable in three ways: (1) loss constant with frequency; (2) loss decreasing with frequency (inversely to the cable); and (3) "bulge" or "bow" up or down in the middle frequency range. By properly selecting the type of network and the amount of loss, it has been possible to equalize a wide variety of circuits. The general shape of the frequency characteristics necessary for a typical line are shown in Fig. 9. The over-all transmission characteristics of video circuits of this kind are shown in Fig. 10.

Special Cables

Special cables with which we have experimented for television transmission have been designed on a balanced shielded basis but are otherwise similar to coaxial cables. They make it possible to span distances of 5 miles or more between amplifiers. The same type of amplifiers and equalizers may be used as indicated above for ordinary telephone cables. Measurements on a special cable containing shielded balanced pairs installed in New York City and equipped with amplifiers indicate that there are no serious technical difficulties

involved in providing such facilities. A typical transmission characteristic is shown on Fig. 10.

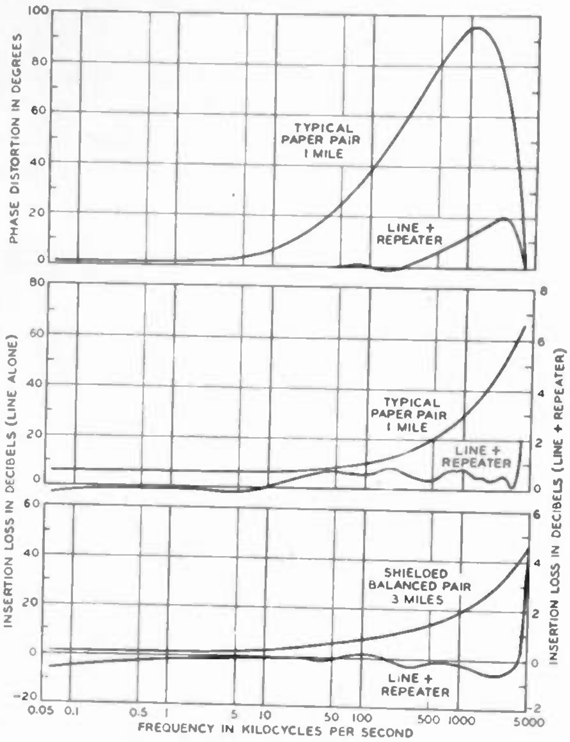


Fig. 10—Over-all characteristics of some video circuits.

Republican Convention Broadcast

In June, 1940, various scenes from the Republican convention in Philadelphia were televised, transmitted to New York over wire lines, and broadcast from the Empire State Building. This experiment in network broadcasting was made jointly by the National Broadcasting Company and by the American Telephone and Telegraph and associated companies. The wire-line circuits consisted of about 96 miles of coaxial cable and about 9 miles of ordinary paper-insulated telephone cable. A total of about 50 amplifiers were employed. The over-all transmission characteristics of this circuit are given in Fig. 11. As shown,

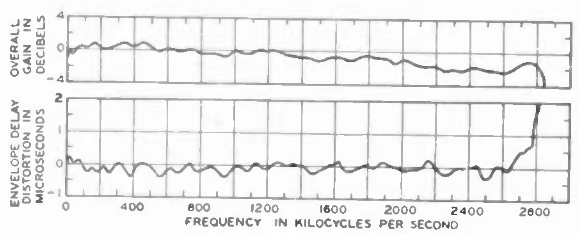


Fig. 11—Republican convention circuit over-all characteristic.

the attenuation is constant over the frequency band within about ± 1.5 decibels and the delay distortion is within about ± 0.3 microsecond.

The over-all characteristics of the circuit demonstrated have about the same transmission characteristics in spite of the fact that the total length of circuit is nearly twice as great.

Methods of Measurement

It is of prime importance in setting up any transmission circuit to know the characteristics of the medium and the apparatus which goes with it to send and receive the signal. Since the television signal is essentially a transient phenomenon it would be logical to assume that a method of measuring the performance of a system which is based on its transient response would be the proper one. There are certain difficulties, however, in interpreting the cause of some of the transients resulting from distortions in the system. After such defects have been located and corrected the final criterion of the result must, of course, be the transient picture itself. For fundamental studies of distortions in the

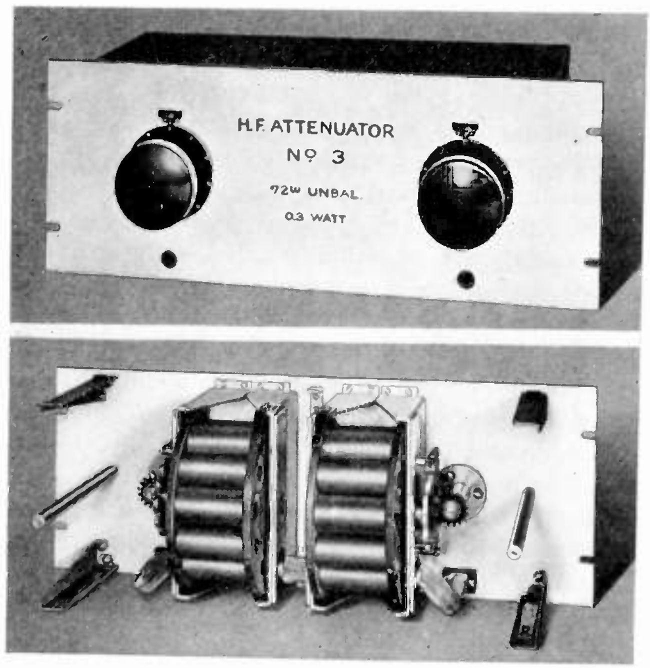


Fig. 12—Photograph of coaxial attenuator.

parts of a system or for locating their position in the combined system, we have continued to use, mostly, the more laborious steady-state technique. In some cases testing with pulses and square-top signals has been useful in determining what particular characteristic to study in greater detail. The necessary accuracy has always been obtained by the steady-state methods.

For line studies it is necessary to measure the attenuation and phase constants before and after equalization. In the design of new lines the distributed constants are most important. The impedance of elements which go into the apparatus attached to the line and the impedance of the lines themselves must also be measured. This section reports the methods we have used and the order of accuracy we have obtained.

Insertion Loss, Attenuation

Single-frequency measurements of insertion loss in the range from 50 kilocycles to 4 megacycles are made with a calibrated oscillator and detector which must

be stable during the time of a single measurement. The reference standard is an adjustable attenuator, accurate to 0.1 decibel and with negligible phase shift up to 4 megacycles.

For checking the working standards a primary laboratory standard of the "coaxial" type is used. The construction is shown in Fig. 12. The units in this attenuator are unusual in that the series elements of the pads are plated rods used as central conductors in coaxial outer shells. The shunt elements are plated disks. Due to the symmetry of construction and the thin conducting surfaces, there is no skin effect in the resistances and only the theoretical phase shift for a few inches of coaxial structure. The units have been adjusted to be better than 0.1 decibel for values up to 40 decibels. Its loss has no measurable change with frequency up to 100 megacycles and it is probably constant well beyond this frequency.

Where the over-all loss is not great, as is the case with repeaters in place, straightaway measurements of sent and received voltage are made with two high-frequency thermocouples and a calibrated oscillator. Fig. 13 shows a field thermocouple set. The range of loss measurable is limited to about 10 decibels from 1 milliwatt for an accuracy of 0.1 decibel.

Phase and Delay

In an over-all television system, as explained above, if the phase shift is proportional to the frequency from the lowest to the highest frequency in the transmitted range, the system will have no phase or delay distortion. Departures from this ideal condition generally manifest themselves in the received picture in two ways: (1) reduction in sharpness of edges and (2), ghosts. The type of phase distortion associated with (1) is a gradual departure of phase shift from linearity with the frequency and that associated with (2) is an abrupt change in the phase shift, or a periodic variation of phase shift from linearity with the frequency. Apparatus has been developed with which either type of distortion may be measured with an accuracy several times as good as that required for satisfactory picture transmission.

In the coaxial carrier system the phase distortion is of importance only in the frequency range transmitted over the coaxial. Thus, in the carrier system there is no distortion if, over the entire frequency range transmitted,

$$(\beta_f - \beta_c) = k(f - c)$$

β_f and β_c are the phase shifts at any frequency (f) and at the carrier frequency (c), respectively, and (k) is some constant.

On over-all television systems the phase shift of single frequencies has generally been measured up to about 50 kilocycles. At higher frequencies, envelope-delay measurements have been made, either on a loop or straightaway. For loop measurements, these envelope-delay measurements give

$$(\beta_{f+\Delta} - \beta_f)/2\pi\Delta$$

delay in microseconds where β is expressed in radians and Δ is the frequency interval in megacycles between the two frequencies used for the measuring envelope. In the case of straightaway measurements, the result is

$$(\beta_{f+\Delta} - \beta_f)/2\pi\Delta + K$$

in microseconds, where K is a constant which does not affect the distortion. In either case the actual phase departure from linearity with frequency is obtained by the summation of the envelope-delay measurements.

In order to determine the performance of a television system we have usually obtained the gradual departure of phase shift from linearity with frequency by means

Fig. 13—Photograph of thermocouple set.

of envelope-delay measurements using a wide frequency interval (perhaps 50 kilocycles). If there are abrupt or short-period departures from linearity, these wide-interval measurements are supplemented by envelope-delay measurements using a narrower interval (perhaps 8 kilocycles). In the case of the New York-Philadelphia coaxial loop, measurements with a 50-kilocycle interval were sufficient to determine the phase-equalizer design, the short-period variations being small.

Loop phase measurements at frequencies below 5 kilocycles have generally been made by introducing a variable phase shift in a reference circuit and making a bridge balance of the received signal from the loop and the signal from the reference circuit. An accuracy of better than 0.1 degree has usually been obtainable by this method, the limit being set by noise in the system being measured. At frequencies above 5 kilocycles, the test frequency has usually been modulated to 50 kilocycles so as to use one 360-degree phase shifter for all

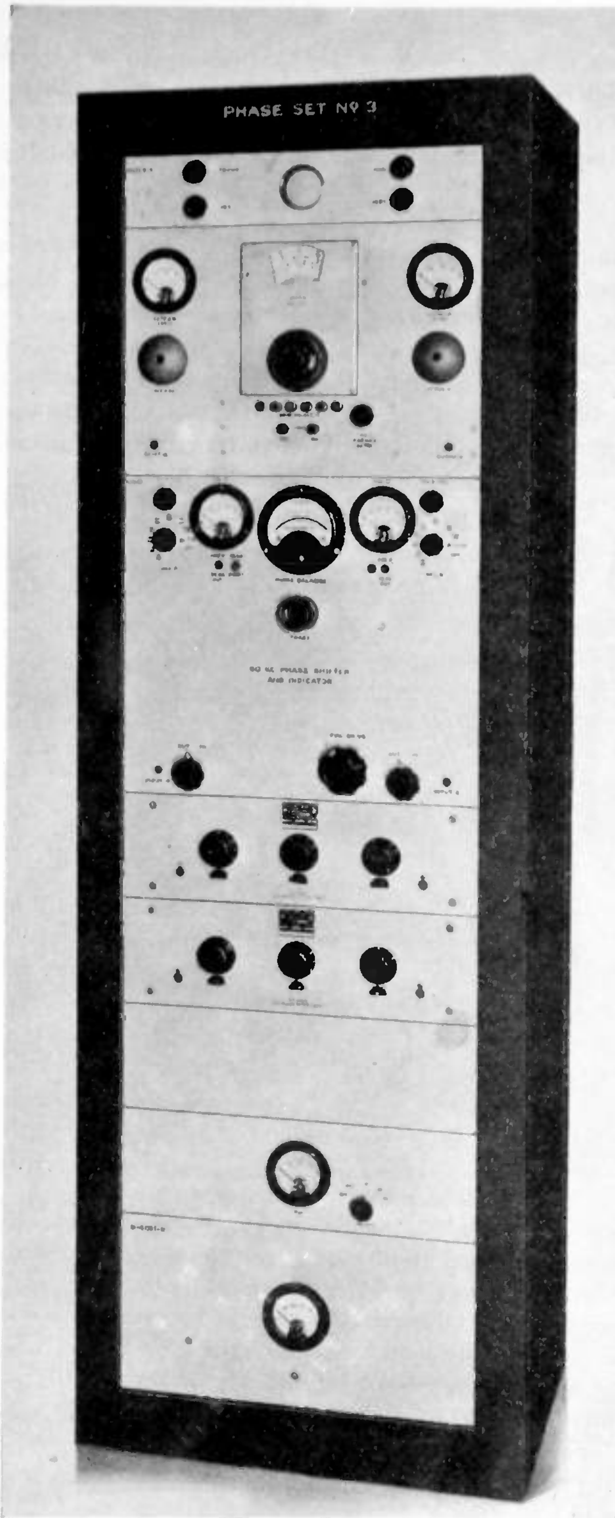


Fig. 14—Photograph of phase-measuring set No. 3.

measurements. With this method the accuracy has generally been better than 0.1 degree. Fig. 14 shows such a phase-measuring set and Fig. 15 the component parts.

A schematic diagram for a straightaway envelope-delay distortion-measuring set is shown in Fig. 16. The two envelope frequencies are produced either by a pair of oscillators held to a definite frequency difference by an automatic-frequency-control circuit as shown or by a balanced modulator. At both the sending and receiving ends the interval frequency is obtained by demodulation, and the phase difference (plus a constant phase shift due to the return circuit and associated equipment) between the outputs of the two demodulators determined by means of the phase

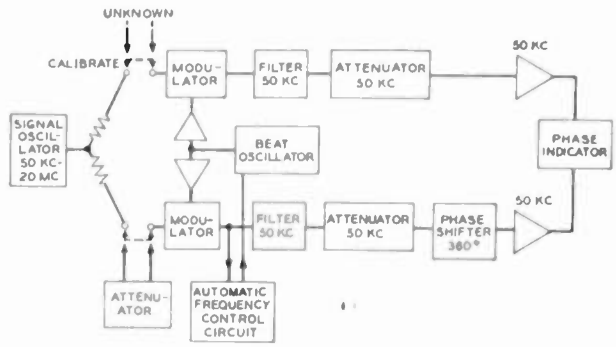


Fig. 15—Schematic of phase-measuring set.

shifter. In such straightaway measurements any difference between the two demodulators must be determined with all of the equipment at the same point; but it has been found that this calibration remains unchanged for long periods of time. Where loop measurements are involved the return circuit disappears, and the calibration is obtained by substituting an attenuator for the loop.

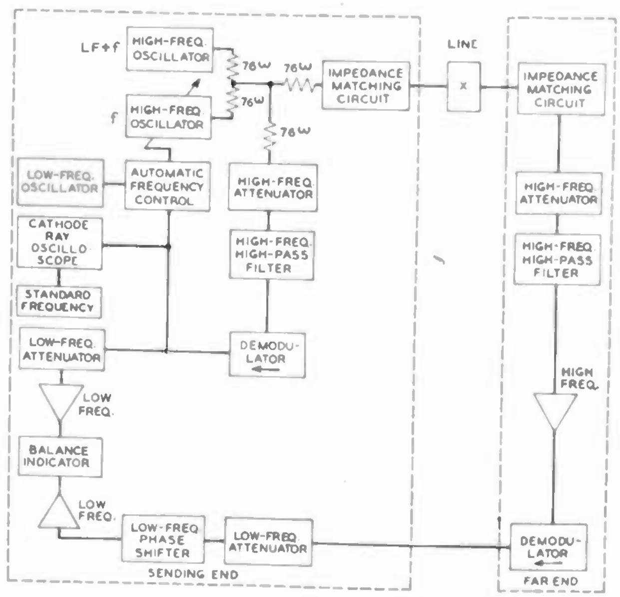


Fig. 16—Block diagram of envelope-delay set.

The accuracy of such delay measurements depends upon the interval frequency Δ. It is about 0.005 microsecond for a 50-kilocycle interval and 0.05 microsecond for an 8-kilocycle interval. For the Republican convention the phase and delay distortions were measured by the above methods. The over-all characteristic including the insertion loss was shown above in Fig. 11.

A further refinement in straightaway delay measurements which is useful on very long circuits is to transmit a reference envelope in the same direction as the test envelope over the unknown circuit and measure the difference in delay for the two envelopes. This gives delay distortion directly. Special filters are required to separate the two envelopes at the far end, but the phase comparison is made in the same way as if the reference interval frequency had been transmitted directly.

Impedance

The characteristic impedance of long lengths of line usually has the irregular appearance of Fig. 3. Here we see the effect of the reflections which occur in the coaxial cable due to various irregularities such as splices or variations in distributed constants. The figure shows the cumulative effect of these echoes on single-frequency values of the input impedance. With a standard high-frequency bridge using variable standards such measurements can be made with an accuracy of ± 0.02 ohm.

In some cases it is more important to know the location of the individual irregularities in the cable and their individual magnitudes. Of great assistance in this problem is a device we have called an "echo" indicator,[14] a simplified diagram of which is shown on Fig. 17. It shows on an oscillograph the reflections produced along the cable due to various irregularities which re-

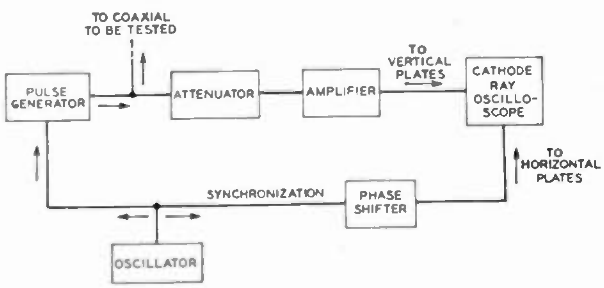


Fig. 17—Schematic of echo indicator.

turn to the sending end. The signal voltage is applied in a form of a sharp pulse about 0.2 microsecond wide, which appears along the vertical axis. A sweep frequency, synchronized with the pulse generator, supplies a horizontal time axis. Let us assume that the pulse is applied to the coaxial unit. When it reaches an irregularity in the cable a fraction of this pulse will be reflected back and will appear as a vertical displacement on the oscilloscope. The time between the application of the pulse and the return of the reflected portion, which will appear as a horizontal displacement on the oscilloscope, is a measure of the distance to the irregularity from the input end of the cable. From the shape and the size of this reflected pulse, or echo, it is possible to estimate certain characteristics of the irregularity, and, in this manner, determine the cause of the irregularity. An example of a typical record of some near-end echoes obtained with this equipment is shown on Fig. 18.

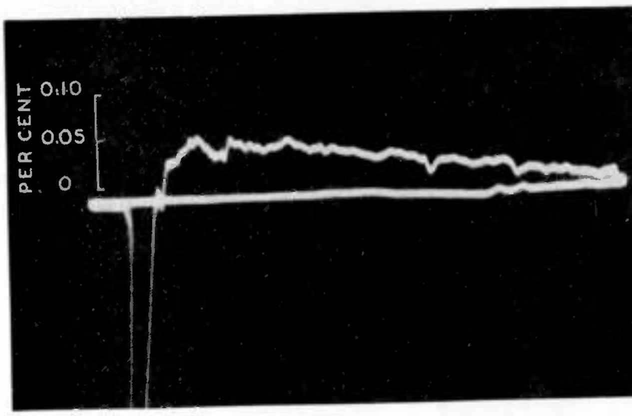


Fig. 18—Near-end echoes in section of coaxial cable.

Conclusion

The experiments reported upon and the accompanying demonstrations which have been made show that wire-line networks of present design are capable of transmitting 441-line, 30-frame interlaced television images over distances of at least some hundreds of miles, without impairment visible to many careful observers. The experiments and theoretical work reported on also show that most of the technical difficulties in such networks increase with length. The various requirements which must be met tend to increase more or less as the square root of the number of repeater sections. The experiments were based on a system which reproduced a rather wide linear brightness range.

The more common defects in present-day television images, as reproduced on commercial receivers, have been minimized in this system by careful transmission design. The design technique has resulted in very small attentuation and delay distortion over the transmitted frequency band. The development of means of measuring accurately the distortions which were found was an important factor in achieving this result.

Measurement of Loop-Antenna Receivers*

W. O. SWINYARD†, MEMBER, I.R.E.

Summary—*The most practical method of introducing the measuring signal into loop receivers is to employ a transmitting loop connected to the signal generator. Several such loops are described, varying with regard to the use of shielding, frequency range covered, method of computing the field strength produced, and other particulars. The types described in the latter part of the paper are in use and giving satisfactory service at the present time. Precautions against stray fields and near-by metal objects are discussed briefly. Measurements on receiving loops, aside from general receiver tests, are described and formulas given. The results of such measurements on three groups of loops, totaling 30 in number, are reported. The paper closes with an Appendix giving the derivation of a few formulas of interest.*

THE widespread adoption of tuned loops as signal pickup devices on broadcast receivers has placed on engineers the problem of devising performance measurements adapted to this type of input system. Suitable methods for introducing and measuring the input radio-frequency voltage to the loop are required.

Two Methods of the Institute of Radio Engineers Standards

In the Institute of Radio Engineers "Standards on Radio Receivers: 1938" there are two alternative methods for introducing a signal into a loop for measurement purposes. One method calls for opening the loop circuit and inserting a resistance across which a measured voltage is developed. This involves some inconvenient procedures. A physical break must be made in one of the loop connections for the insertion of the resistor, the constants of the loop must be known or measured in order to reduce the data to equivalent field strength, and the effect of neighboring conductive or ferromagnetic bodies on the field distribution is difficult to evaluate. In addition, the calculations involved are somewhat laborious. The time required for these steps makes the measurements slow.

The other method is very simple. It merely involves the use of a transmitting loop to create an induction field of known strength into which the receiving loop, or the receiver as a unit, is placed. The measurements yield results which are already expressed in terms of field strength, that is, in microvolts per meter. Therefore, apparatus for making measurements on loop receivers is generally based on this method.

Multiturn Loop Operated without Series Resistance

One of the best types of transmitting loops in use today is an outgrowth of several previously used types. One of these earlier types consisted of 19 turns of No. 18 double-cotton-covered copper wire close wound on a bakelite tube having an outside diameter of 3 inches.

* Decimal classification: R261×R325.3. Presented, Chicago section, December 1, 1939. Reprinted from the RMA Technical Bulletin, December 18, 1940, by permission of the Radio Manufacturers Association, Washington, D. C.

† Hazeltine Service Corporation, Chicago, Ill.

This was connected to the output of the signal generator through a low-capacitance shielded transmission line. The generator output terminals were carefully shielded and the ground side of the transmitter loop was faced toward the receiver loop to reduce the capacitive transfer from the generator to the receiver loop. The loops were set up coaxially, and the separation between the centers was made 24 inches in order to maintain a ratio of at least 2 to 1 between this distance and the largest dimension of the usual receiving loop.

For quantitative results, attention may first be directed to equation (1) of the I.R.E. Standards, page 25, which can be written as follows:

$$\mathcal{E} = \frac{18.85 N r^2 I}{X^3},$$

where $\mathcal{E}$ = field strength at the receiving loop in microvolts per meter
N = number of turns in transmitting loop
r = radius of transmitting loop in centimeters
I = current through transmitting loop in milliamperes, and
X = distance between centers of transmitting and receiving loops in meters.

It is convenient to express the current in terms of the voltage, inductance, and frequency, and to express most of the lengths in inches. The formula then becomes

$$\mathcal{E} = \frac{1180 N r^2 E}{X^3 f L}$$

where $\mathcal{E}$ = field strength of receiving loop in microvolts per meter
N = number of turns in transmitting loop
r = radius of transmitting loop in inches
E = microvolts across transmitting loop
X = distance between centers of transmitting and receiving loops in inches
f = frequency in megacycles, and
L = inductance of transmitting loop in microhenrys.

In the present case the inductance of the 19-turn coil on the 3-inch form was just 36.5 microhenrys so that the formula for the 24-inch distance simplifies to

$$\mathcal{E} = \frac{E}{10f}.$$

This makes the calculation quite simple, since the field strength is obtained by dividing the signal-generator output reading by 10 times the frequency in megacycles.

The impedance of the transmitting loop at the low-

frequency end of the broadcast band is greater than 100 ohms so that it can be shunted across the usual 5-ohm resistive attenuator without causing appreciable loss of voltage.

Unshielded Loops with Series Resistance

Another type of transmitting loop, which represents a considerable improvement over the one just described will now be considered. This loop has a 10-centimeter radius and consists of two turns of quarter-inch copper tubing spaced 1/4 inch. This loop has a very low reactance, about 16 ohms at 1700 kilocycles, and hence the electrostatic coupling effects are small. However, the low-potential face is turned toward the receiver loop in order to secure further reduction of such effects. Since it is desirable to eliminate the frequency factor from the formula, the inductance is made small; then, in order to maintain a suitable impedance for the load across the signal generator, a series resistance is inserted in the loop circuit. The resistor is proportioned to make the signal strength in microvolts per meter at the chosen distance of 50 centimeters equal to one tenth of the signal-generator reading in microvolts.

The value of this resistance will now be obtained. It will be satisfactory to have a field strength of 100,000 microvolts per meter for 1 volt across the terminals of the transmitting loop. Set $\mathcal{E}$ equal to 100,000 microvolts per meter in the equation from the I.R.E. Standards, substitute the known values of N, r, and X, and solve for I. The result is that the current must be 3.31 milliamperes. Since this is for a 1-volt potential, the resistor must be 302 ohms. A simple calculation shows that the reactance of the loop has a negligible effect on its impedance.

The resistor, the transmission line to the generator, and the generator output terminals are completely shielded, but the loop itself is not shielded.

Details of another transmitting loop similar to this one are given in Supplement A to the Radio Manufacturers Association Instruction Book 15-38C. This loop consists of 3 turns of No. 20 wire wound 6 turns per inch on a form having a 20-centimeter diameter; it is used at a distance of 50 centimeters from the receiving loop. In this case the series resistance is 450 ohms.

Shielded Loop with Series Resistance

An additional type of transmitting loop to be considered is completely shielded electrostatically. A diagram of it is given in Fig. 1, and a view is given in Fig. 2. This loop is 10 inches in diameter and consists of three turns of No. 20 solid tinned copper wire with celanese insulation. These turns are placed in a copper tube which is bent into a circular form having ten inches as the mean diameter. The copper tube is prevented from acting as a short-circuiting turn by attaching one end only to ground and insulating the other end. Through the gap between the two ends of the copper tube are brought the start and finish connections of the loop. A small housing at the base of the transmitting loop contains a 403-ohm resistor which is connected in series between the ungrounded end of the loop and the high side of the shielded cable leading

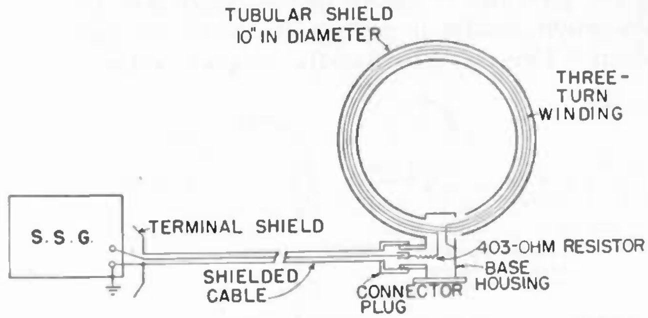


Fig. 1—Diagram of 3-turn shielded loop with 403 ohms in series.

to the standard signal generator. A quarter-inch shielded microphone cable, 4 feet long, is used; this terminates at the loop end in a single-connection-and-ground microphone plug, which screws into a jack in the base of the transmitter loop. At the other end the cable is connected to the ground and high sides of the signal generator by a plug combined with a metal shield sufficiently large to prevent capacitive coupling between the generator output terminals and the receiver loop.

For convenience of manipulation the loop and housing are designed to be mounted on a regular microphone stand.

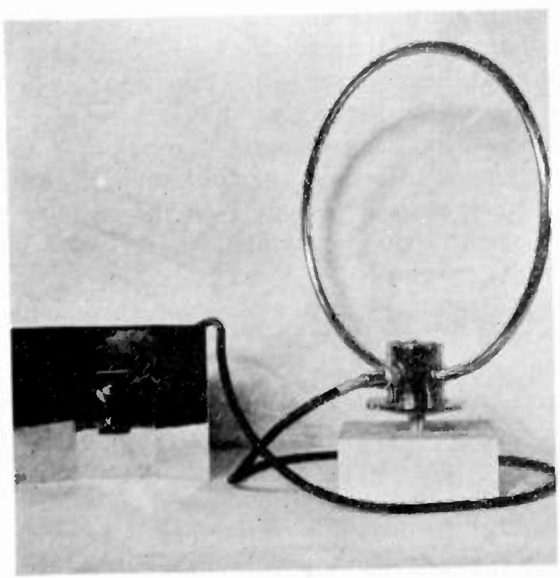


Fig. 2—View of 3-turn shielded loop.

The value of 403 ohms was determined by the method already described; it applies for a distance of 24 inches between the transmitting and receiving loops. With these conditions a field strength equal to one tenth the generator reading exists at the receiving loop.

It should be pointed out that the equation in the Institute of Radio Engineers Standards introduces an

error of about 6 per cent for a transmitting loop with a 10-inch diameter used at a distance of 24 inches. This is because the radius of the loop r was assumed to be appreciably smaller than the separation X permitting r to be dropped from the denominator in the derivation of the formula, as shown in the Appendix. This simplification results in values of sensitivity which are about 0.5 decibel less than the accurate value.

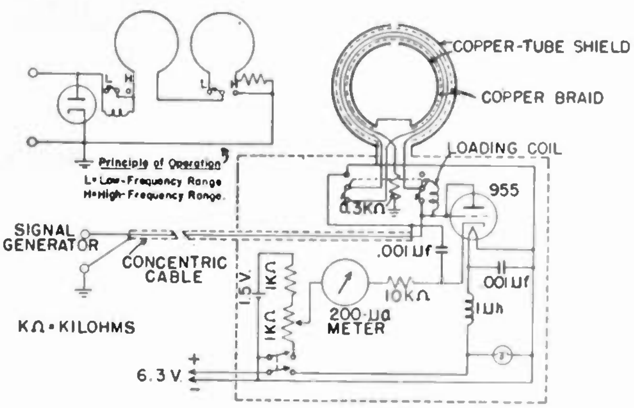


Fig. 3—Circuit diagram of all-wave loop.

It is to be noted that if the output impedance of the signal generator is high, the loading effect of the resistor upon the generator must be taken into account in determining the voltage across the coil, and hence the field strength.

Where broadcast receivers having loops with a dimension exceeding 12 inches are to be measured, the transmitting loop should be separated from the receiving loop by a distance which is at least twice the maximum dimension of the receiving loop. The field strength produced at the receiving loop is inversely proportional to the cube of the distance, so that at 3 feet separation (suitable where the maximum dimension of the receiving loop is 18 inches) the field strength in microvolts per meter is no longer 1/10 the output of the signal generator in microvolts, but is 1/33.7 of the microvolts output.

The maximum frequency at which this equipment can be used is limited by the shunt capacitance in the shielded connecting cable and the series inductance of the cable and loop. Measurements have indicated that with the cable length of 4 feet and a constant voltage input to the line, the rise in voltage at the end of the cable, as the frequency is increased, is sufficiently balanced by the drop due to the increasing reactance of the loop to provide nearly uniform current through the loop up to a frequency of 20 megacycles. The capacitance of the 4-foot length of cable is approximately 120 micromicrofarads, and the series inductance of the cable and loop is approximately 7.5 microhenrys.

Two-Band Shielded Loop

Another type of transmitting loop was specifically designed to provide for accurate measurements of loop receivers at all frequencies from 0.1 to 24 megacycles. A diagram of this is given in Fig. 3 and a view in Fig. 4. This is a 2-band design; on the lower band from 0.1 to 5 megacycles the loop and operation are the same as in the type just described except that the dimensions and number of turns are different. Two turns, 8 inches in diameter, in series with a 300-ohm resistor are used. The field strength in microvolts per meter for a separation of $19\frac{7}{8}$ inches is obtained by dividing the input voltage to the loop by 10.

Measurements on the higher band from 5 to 24 megacycles are made by using only one of the turns and having a loading coil in series to make the total series inductance of the loop circuit have the value of 2.45 microhenrys. The natural frequency of this loop is well above the band, so that appreciable errors are not introduced from this source. For this range the field strength in microvolts per meter, when the loops are spaced $19\frac{7}{8}$ inches coaxially, is equal to the voltage input to the loop divided by the frequency in megacycles.

A diode tube voltmeter is built into a small housing at the base of the loop so that the input can be measured directly at the loop terminals, thus avoiding any uncertainty due to losses or resonant effects in the transmission line.

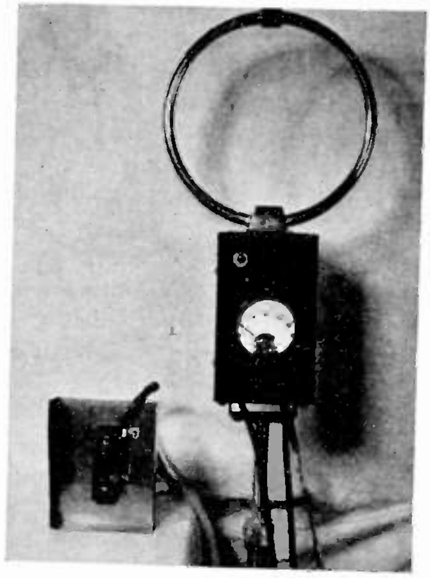


Fig. 4—View of all-wave loop.

If measurements are made using a generator with an attenuator calibrated in decibels and if the results are to be expressed in decibels, the field strength in the range from 5 to 24 megacycles is simply the attenuator reading in decibels plus the decibels corresponding to the frequency in megacycles. The figures are added instead of subtracted because both are taken as positive and an increase of either one means a more sensitive receiver.

For testing receivers with loops having a dimension in excess of 10 inches, the separation between the loops

should be increased. If this is made $42\frac{5}{8}$ inches, the conversion formulas become

$$\mathcal{E} = \frac{E}{100} \text{ for the range of 0.1-5 Mc. and.}$$

$$\mathcal{E} = \frac{E}{10f} \text{ for the range of 5-24 Mc.}$$

The loop is double-shielded to minimize extraneous coupling due to current induced in the shielding.

Although some radiation from the transmission line has been encountered at high frequencies, it is sufficiently low in magnitude for its effect to be neglected if reasonable separation is maintained between it and the loop under test.

Error Due to Leakage Field

In making measurements on sensitive loop receivers trouble due to generator leakage is often experienced. It is necessary that this be minimized by proper orientation of the receiver or generator, or by further shielding of the generator, if accurate results are to be secured. The presence of leakage fields can readily be detected by turning the transmitting loop through 180 degrees. If this produces appreciable change in the measured sensitivity, the cause is likely to be leakage of some type.

One other precaution should be mentioned. It is important to keep both the transmitting and receiving loops well away from any large metal objects which would distort the magnetic field. This applies to the conducting walls of shielded rooms. A clearance around each loop equal to twice the distance between loops is considered a satisfactory minimum value. The receiving loop is generally left in place in the receiver cabinet because the effect of the conducting chassis is desired since it is present in normal operation.

Measurements on the Loop Only

In addition to measurements of over-all sensitivity there are several tests generally made on the loop itself. One of these is a determination of the loop Q from band-width measurements made on the loop when it is in its proper place in the cabinet; this is usually done at 3 frequencies. These band widths are measured with a large signal in the transmitting loop and with the 2 loops separated as far as possible while still affording an adequate reading at resonance on a tube voltmeter. This procedure avoids appreciable loading of the receiving loop under test. The relationship between the 2 times band width W_2 and Q is

$$Q = \sqrt{3}\left(\frac{f}{W_2}\right),$$

where W_2 and the frequency f are in the same units.

The distributed capacitance is calculated from the formula,

$$C = \frac{C_1 - 4C_2}{3},$$

where C_1 is the capacitance for tuning to any frequency f_1 and C_2 is the capacitance for tuning to $2f_1$. These values can be conveniently measured on a reactance meter.

The loop figure of merit, which is defined as Q times the effective height, is obtained by dividing the microvolt sensitivity of the receiver on the first grid by the over-all sensitivity in microvolts per meter. These measurements are usually made at 3 points in the band; they are accurate only when the receiver is aligned at these frequencies and no regeneration is present.

Characteristics of Loops Used in Recent Receivers

Measurements have been made on 30 loops of various kinds and sizes, supplied by coil or radio manufacturers or made in our laboratory. All determinations were made with the loop in free space. The values of the distributed capacitance, of Q, and of the figure of merit were measured for each loop. The loops were then arbitrarily divided into three groups according to area: small, 20 to 35 square inches; medium, 36 to 50 square inches; and large, over 50 square inches. Averages of the measurements were then computed for each group, with the results shown in Table I.

An attempt was made to establish empirically a means of determining the effective area from the maximum and minimum areas, so that calculated values of the figure of merit (which is Q times the effective height or $Q \cdot 2\pi NA/\lambda$) would be in closer agreement with measurements. However, no consistent relationship could be found.

Experience has shown that in most cases loops are mounted so close to the receiver that the Q and figure of merit are materially reduced below the values in Table I, so that these figures indicate better performance than is usually obtained.

TABLE I
Average Characteristics of Three Groups of Loops

Type Loop	Number in Each Group	Average Area, square inch	Average Number Turns	Average Capacitance, micromicrofarads	Average Q			Average Figure of Merit		
					600 kilocycles	1000 kilocycles	1400 kilocycles	600 kilocycles	1000 kilocycles	1400 kilocycles
Small	7	29	35	7.4	117	111	93	0.52	0.73	0.84
Medium	11	42	28	11.0	126	118	92	0.75	0.99	1.00
Large	12	76	22	13.0	140	130	95	1.50	2.00	2.13

Acknowledgment

The measuring loops described in this paper are the results of work by various engineers of the Hazeltine Service Corporation. Acknowledgment is particularly due to N. P. Case of the New York Laboratory and J. Kelly Johnson of the Chicago Laboratory.

Appendix

This section is devoted to the derivation of a few formulas which are believed to be of general interest to those concerned with the design of transmitting loops used for making measurements on loop receivers.

The first derivation is that for equation (1), page 25, given in the Institute of Radio Engineers "Standards on Radio Receivers: 1938." Refer to Fig. 5.

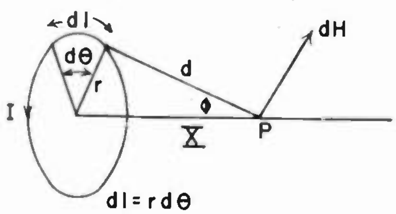


Fig. 5—Derivation of formula for induction field.

Using rationalized mks units,[1]

$$dH = \frac{Idl}{4\pi d^2} = \frac{Ir d\theta}{4\pi d^2}.$$

$$dH_x = (\sin\Phi)(dH) = \frac{r}{d}\,dH = \frac{Ir^2 d\theta}{4\pi d^3}.$$

Integrating,

$$H = H_x = \frac{Ir^2}{4\pi d^3}\int_0^{2\pi N} d\theta,$$

where N = number of turns.

Performing the integration,

$$H = \frac{Ir^2N}{2d^3}.$$

Introducing X,

$$H = \frac{Nr^2I}{2(X^2+r^2)^{3/2}}.$$

Now the field strength in volts per meter is[2]

$$\mathcal{E} = 120\pi H$$

whence

$$\mathcal{E} = \frac{120\pi\cdot Nr^2I}{2(X^2+r^2)^{3/2}} = \frac{188.5Nr^2I}{(X^2+r^2)^{3/2}}.$$

[1] Harnwell, "Principles of Electricity and Magnetism," pp. 286–289 and 600–605.
[2] *Ibid.*, p. 535.

Assuming that $r^2 \ll X^2$,

$$\mathcal{E} = \frac{188.5Nr^2I}{X^3}.$$

This is in volts, meters, and amperes. Changing to microvolts per meter for $\mathcal{E}$, centimeters for r, and milliamperes for I, leaving X in meters, there is obtained

$$\mathcal{E} = \frac{18.85Nr^2I}{X^3} \quad \text{or} \quad 4\mathcal{E} = \frac{75.4Nr^2I}{X^3},$$

which is the expression in the Institute of Radio Engineers Standards.

Another matter of interest is the voltage induced in a receiving loop by a given transmitting loop, or in a given field. Referring to Fig. 6, the derivation of a

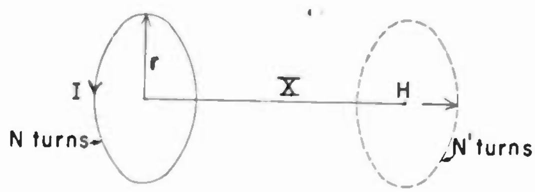


Fig. 6—Derivation of formula for voltage induced in receiving loop by a transmitting loop.

general formula for any two loops, followed by an application for any receiving loop at the standard distance from the 3-turn shielded loop of Figs. 1 and 2 is:

Using rationalized mks units, as in the preceding figure,

$$H = \frac{Ir^2N}{2X^3}.$$

Substituting $r^2 = A/\pi$, where A is the area of the transmitting loop,

$$H = \frac{IAN}{2\pi X^3}.$$

The mutual inductance between the two loops placed coaxially is

$$M = \frac{\mu_0 HA'N'}{I}.$$

Substituting for H,

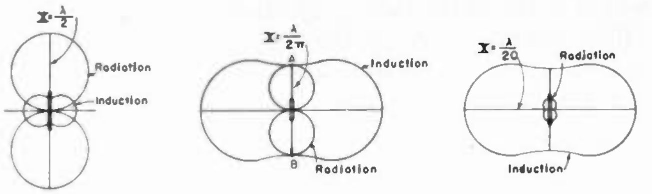
$$M = \frac{\mu_0 A'N'}{I}\left(\frac{IAN}{2\pi X^3}\right) = \frac{\mu_0 AA'NN'}{2\pi X^3}.$$

The desired voltage in the receiving loop is

$$E_2 = I\omega M = \frac{I\omega\mu_0 AA'NN'}{2\pi X^3}.$$

Substituting $\mu_0 = 4\pi\times10^{-7}$, the permeability of space, and replacing ω by $2\pi f$, there is obtained

$$E_2 = 4\pi\times10^{-7} IfAA'NN'/X^3.$$



Type of Antenna	Radiation Field	Induction Field		Distance at Which Fields are Equal
		In Plane of Loop	Perpendicular to Plane of Loop	
Loop	$\frac{\pi IAN}{\lambda^2 X}$	$\frac{IAN}{4\pi X^3}$	$\frac{IAN}{2\pi X^3}$	$\frac{\lambda}{2\pi}$
Vertical	$\frac{hI}{2\lambda X}$	$\frac{hI}{4\pi X^2}$		$\frac{\lambda}{2\pi}$

Field strengths are in ampere turns per meter, currents in amperes, all lengths in meters.

Fig. 7—Induction and radiation fields of loop and vertical antennas.

This is the desired relation in volts, cycles per second, and meters.

Applying to the loop of Figs. 1 and 2, this becomes

$$E_2 = 2.08 \times 10^{-8} \mathcal{E} f A' N'.$$

Converting for kilocycles and square inches, leaving $\mathcal{E}$ in microvolts per meter,

$$E_2 = 1.35 \times 10^{-8} \mathcal{E} f A' N'.$$

Fig. 7 shows graphically the relationship between the induction and radiation fields about a loop, and gives formulas by which these fields may be computed. A comparison of the fields surrounding a loop and a vertical antenna is also made. The middle diagram shows at points A and B the equality of the two fields at a distance of $\lambda/(2\pi)$ which is the basis of the choice of this distance as the limit for 15 microvolts per meter by the Federal Communications Commission for short-distance wireless phonograph and remote-control apparatus.

Errors in the Calibrated Losses of Symmetrical Resistance Networks*

ARTHUR W. MELLOH†, ASSOCIATE, I.R.E.

Summary—*Symmetrical resistive attenuating networks are usually designed to produce a definite calibrated transmission or insertion loss when working between given terminal resistances. When the terminations differ from the value assumed in the design, the transmission and insertion losses caused by the network will no longer be equal to each other nor to the calibrated loss, and this paper discusses the corrections that must be applied to obtain the true losses. Equations and curves are given from which the corrections may be obtained for cases of practical importance.*

Introduction

THE METHODS for designing a 4-terminal resistance network to satisfy given attenuation and impedance requirements are well known, and extensive numerical tabulations[1,2] permit the selection of suitable network elements with a minimum of calculation for a wide range of conditions. It is not always possible, however, to provide a network to suit each new situation and in practical cases it is often necessary to use a fixed or adjustable attenuator under conditions other than those for which it was designed. If, for example, a resistance network or pad which has been designed to work between 600-ohm terminations and cause a loss of 5 decibels is connected between 500-ohm terminations, the loss introduced by it is no longer 5 decibels and it is desirable to know what correction must be made to obtain the true loss. The problem is further complicated by the fact that either the transmission loss or the insertion loss may be desired, depending on the viewpoint taken and the type of measurement being made. In the example just mentioned, each of these losses will be equal to the calibrated loss of 5 decibels only when the network is terminated in 600 ohms, and for other terminations the transmission and insertion losses will not only differ from 5 decibels but will also differ from each other. Relatively large errors may result if these differences are not taken into account; and, even in cases where the errors are negligible, a knowledge of their existence and possible magnitude usually leads to a more intelligent use of the attenuator.

The analysis will be confined to symmetrical net-

* Decimal classification: R383. Original manuscript received by the Institute, March 31, 1941.

† Associated Electric Laboratories, Inc., Chicago, Ill.

[1] P. K. McElroy, "Designing resistive attenuating networks," Proc. I.R.E., vol. 23, pp. 213–233; March, 1935.

[2] C. D. Colchester and M. W. Gough, "Resistance networks," *Wireless Eng.*, vol. 17, pp. 206–215; May, 1940.

works since most adjustable attenuators are of this type. Fig. 1 shows such a network, reduced to its equivalent T form, which receives energy from a generator having an internal electromotive force of E volts and internal resistance of R_1 ohms, and delivers energy to a load resistance of R_2 ohms. It is assumed

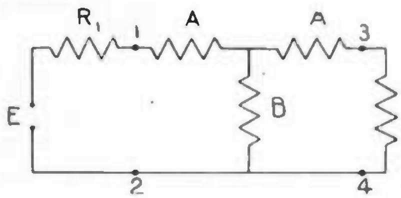


Fig. 1—Symmetrical 4-terminal resistance network.

that A and B have been determined so that the network has an image resistance of R_K ohms and causes a loss of N_C decibels or α nepers, the relation between N_C and α being

$$N_C = 10 \log_{10} e^{2\alpha}. \tag{1}$$

A convenient and well-known form in which the design equations may be written is

$$\left.\begin{aligned} A &= R_K \tanh \alpha/2 \\ B &= R_K/\sinh \alpha \end{aligned}\right]. \tag{2}$$

The transmission and insertion losses will each be equal to N_C when $R_1 = R_2 = R_K$, and it is interesting to observe that both of these viewpoints have been used[1,2] in discussions of attenuator design. N_C will be referred to as the calibrated loss since it is the value normally associated with a given setting of an adjustable network. This paper discusses the algebraic additions to N_C that must be made to obtain the transmission loss N_T or the insertion loss N_I when either R_1 or R_2, or both, differ from R_K.

Transmission Loss

The transmission loss in decibels through the network of Fig. 1 is defined by

$$N_T = 10 \log_{10} P_1/P_2 \tag{3}$$

where P_1 is the power delivered to the network at terminals 1 and 2 and P_2 is the power delivered to the load at terminals 3 and 4. It is only when R_1 and R_2 are both equal to R_K that $N_T = N_C$, and when the terminal resistances are different from R_K the application of (2) and (3) to Fig. 1 gives

$$N_T = 10 \log_{10} e^{2\alpha} + 10 \log_{10} \frac{(R_K + R_2)^2}{4R_K R_2} + 10 \log_{10} \left[1 - e^{-4\alpha} \left(\frac{R_K - R_2}{R_K + R_2} \right)^2 \right]. \tag{4}$$

Using (1), this becomes

$$N_T - N_C = 10 \log_{10} \frac{(R_K + R_2)^2}{4R_K R_2} + 10 \log_{10} \left[1 - e^{-4\alpha} \left(\frac{R_K - R_2}{R_K + R_2} \right)^2 \right] \tag{5}$$

which is the correction in decibels to be added to the calibrated loss to obtain the true transmission loss. It is observed that N_T is independent of the generator resistance R_1. The correction given by (5) consists of two terms, the first of which is constant for a given network and termination, and is recognized as the expression for the reflection loss between R_K and R_2. The second term contains α and therefore will change with each setting of an adjustable network. Since the sum of the two terms of (5) is positive regardless of whether R_2 is greater or less than R_K, the transmission loss will always be greater than the calibrated loss except when the network is terminated in R_K. From the symmetry of the two terms it is evident that the correction will have the same numerical value for a given ratio of R_K and R_2 as for the reciprocal ratio. Fig. 2 gives the cor-

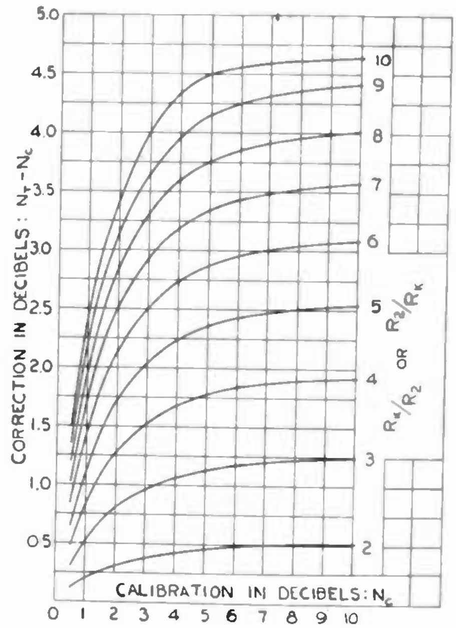


Fig. 2—Correction to calibrated transmission loss.

rection as calculated from (5) for various ratios of R_K and R_2. These curves show that for values of N_C greater than about 10 decibels the correction is practically independent of the calibration and depends only on R_K and R_2; that is, for sufficiently large α the second term of (5) becomes negligible and the correction is just the reflection loss between R_K and R_2. The correction for this case is given in Fig. 3. A useful observation is that an adjustable attenuator will increase the transmission loss by increments equal to the increments in the calibration, regardless of the value of R_2, provided that N_C is at least 10 decibels.

Insertion Loss

The insertion loss in decibels caused by the network of Fig. 1 is defined by

$$N_I = 10 \log_{10} P_3/P_4 \tag{6}$$

where P_3 and P_4 are the power delivered to R_2 before and after the network is inserted, respectively. It has been stated previously that when R_1 and R_2 are equal to R_K in Fig. 1, the insertion loss caused by the network is equal to the calibrated loss and is given by (1). A calculation of the insertion loss for other values of terminal resistances, together with (1) and (2) gives

$$N_I - N_C = 10 \log_{10} \frac{(R_K+R_1)^2}{4R_KR_1} + 10 \log_{10} \frac{(R_K+R_2)^2}{4R_KR_2} - 10 \log_{10} \frac{(R_1+R_2)^2}{4R_1R_2} + 10 \log_{10} \left[1 - e^{-2\alpha}\left(\frac{R_K-R_1}{R_K+R_1}\right)\left(\frac{R_K-R_2}{R_K+R_2}\right)\right]^2. \quad (7)$$

A similar equation is often used[3] for determining the losses in electric wave filters since such calculations are usually made from the viewpoint of insertion loss rather than transmission loss, but a discussion of this equation as applied to purely resistive networks does not appear to have been given. Equation (7), which is the correction in decibels to be added to the calibrated loss to obtain the true insertion loss, is a function of both terminal resistances and α. The first three terms give the reflection losses between R_1, R_2, and R_K and hence are constant for a given network with fixed terminations, while the fourth term involves α and will vary with the setting of an adjustable network.

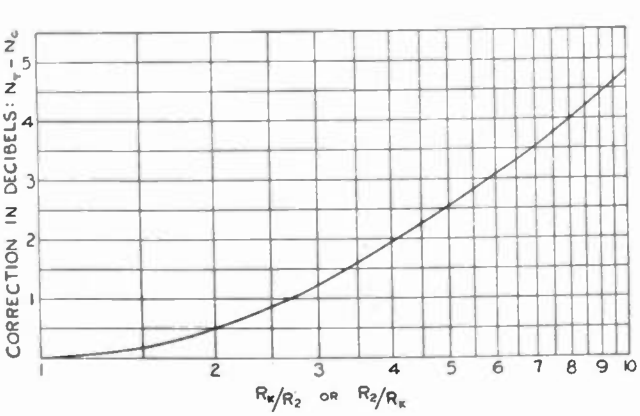


Fig. 3—Correction to calibrated transmission loss for all values of N_C greater than 10 decibels.

Since both R_1 and R_2 may vary independently in practical cases, a large number of curves would be required to portray (7) over a useful range of values. It is often possible, however, to choose the apparatus associated with the network so that the corrections can be determined from a simplification of (7). In many experimental arrangements it is easy to simulate the condition of $R_1 = 0$ by maintaining a constant voltage at the input terminals of the attenuator. For this particular case, (7) reduces to

$$N_I - N_C = 10 \log_{10} \frac{(R_K + R_2)^2}{4R_KR_2} + 10 \log_{10} \frac{R_K}{R_2} + 10 \log_{10} \left[1 - e^{-2\alpha}\left(\frac{R_K - R_2}{R_K + R_2}\right)\right]^2 \quad (8)$$

which is plotted in Fig. 4 for various values of R_K/R_2 and α. It is observed that for values of α greater than 20 decibels the correction is practically independent of

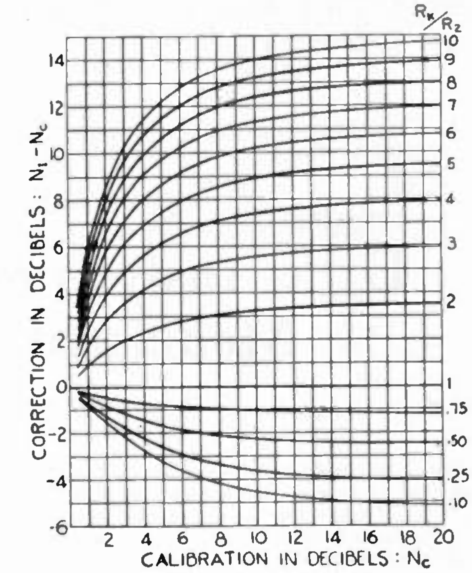


Fig. 4—Correction to calibrated insertion loss when generator resistance is zero.

N_C and is given by the reflection loss between R_K and R_2 plus an additional loss which depends on the ratio R_K/R_2. When the setting of an adjustable network is changed, the insertion loss will be changed by an increment equal to the increment in N_C provided N_C is at least 20 decibels. This value of 20 decibels does not depend on the particular choice of R_1 used for Fig. 4, but holds for any value of R_1, R_2, or R_K. Since the second and third terms of (8) do not take on the same value for the reciprocal of R_K/R_2 as for the ratio itself, it is possible for the correction to be negative. This is true also for the general case expressed by (7), from which it is easy to show that the negative correction can never exceed approximately 6 decibels while the positive correction may assume any value, depending on the values of R_1, R_2, and R_K.

The curves of Fig. 4 are of interest not only because the condition $R_1 = 0$ may often be realized in practice but also because they represent, for the range of R_K/R_2 shown, the maximum possible corrections that may have to be applied for any value of R_1; however, whether the correction is positive or negative for a given value of R_K/R_2 depends on R_K/R_1. This is shown in Fig. 5 where the curves marked $R_K/R_1 = \infty$ and

[3] T. E. Shea, "Transmission Networks and Wave Filters," D. Van Nostrand, Inc., New York, N. Y., 1929, p. 119.

$R_K/R_1 = 0$ represent the maximum possible corrections when R_1 tends to zero and when R_1 increases without limit, respectively. Since these curves give maximum corrections they correspond to the condition of N_C greater than 20 decibels; for smaller values of N_C the corrections will be less than those shown and will vary with N_C. In any case the correction will lie within the area between the two limiting curves and Fig. 5 may be useful for estimating the correction; for greater accuracy, (7) can be used, or Fig. 4 if $R_1 = 0$.

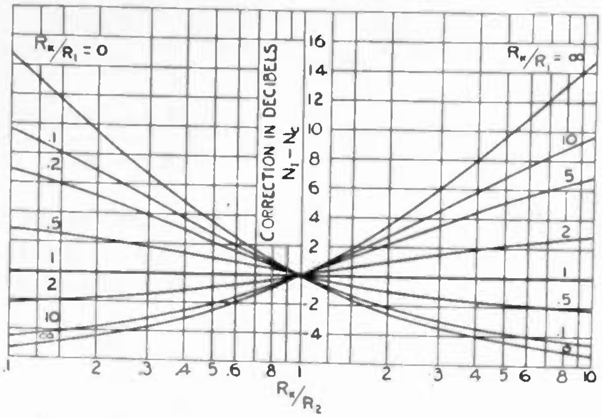


Fig. 5—Correction to calibrated insertion loss for all values of N_C greater than 20 decibels.

An additional important and often unappreciated fact brought out by Fig. 5 is that if either R_1 or R_2 (but not necessarily both) is made equal to R_K, the insertion loss is equal to the calibrated loss for all values of N_C. This, of course, is a condition much to be desired in practical cases since the need for corrections is then eliminated, but circumstances are often such that neither of these conditions can be met and it then becomes necessary to use correction curves or equations such as are presented in this paper.

Drift Analysis of the Crosby Frequency-Modulated Transmitter Circuit*

E. S. WINLUND†, NONMEMBER, I.R.E.

Summary—*Component drift, sensitivity, and band-width expressions are combined in an over-all expression for frequency stability of the Crosby circuit. Using experimentally obtained constants in this expression, an equation is derived for drift in terms of frequency and frequency multiplication open to choice by the designer, and the results are shown as design curves. The equation is checked against actual conditions existing in a Crosby exciter unit.*

Foreword

The recent interest in frequency modulation has lent new impetus to investigations of the many methods of producing it and of their relative stability, fidelity, and simplicity. Because frequency modulation has found its place in the ultrahigh frequencies where absolute stability is inherently difficult to obtain, and because the concept of frequency modulation is directly opposed to that of high stability, stability has been the more formidable hurdle for designers.

Most of the circuits finding general acceptance today require, inevitably, a balance between simplicity, stability, and fidelity, the latter including distortion, noise, and response. Extremely high frequency multiplication introduces random-phase noise, and if the multiplication is obtained by heterodyning with a separate crystal, the frequency stability suffers. Frequency modulation obtained from phase modulation by any of the usual processes inherently limits deviation for acceptable low-frequency distortion, requiring large multiplication.

* Decimal classification: R414. Original manuscript received by the Institute, January 28, 1941. Presented, Sixteenth Annual Convention, New York, N. Y., January 11, 1941.

† RCA Manufacturing Company, Camden, N. J.

Reactance-tube modulators provide a direct frequency-modulating means, without corrective networks. Their distortion is inherently extremely low and independent of the modulating frequency as will be shown in this paper. Their deviation capability may be designed to take care of modulation and superimposed automatic-frequency-control action easily and linearly. The superposition of automatic frequency control necessitates no compromise.

Stabilization, then, may be separately vested in an automatic-frequency-control system, such as that employed in the Crosby circuit, so that over-all stability is a function of the stabilities of the crystal, the modulated oscillator, and the low-frequency discriminator. The advent of inexpensive quartz crystals has rather obscured the stability possibilities of ordinary tuned circuits, especially the *passive* circuits of which the conventional discriminator is a good example. Discriminators can be built to have a stability comparable to that of crystals.

These considerations, together with the simplicity of the Crosby circuit, have led to this analysis, and to a commercial design based upon it. Expressions for the frequency drift in various portions of the Crosby circuit will be presented one at a time and then combined at the end to provide an equation for the drift of the entire Crosby circuit.

Reactance Modulators

The sensitivity and band width of the conventional reactance-tube modulator have been rather simply cal-

culated by N. I. Korman,[1] by assuming a linear transconductance characteristic from cutoff to zero bias. In Fig. 1, the following symbols and definitions will apply, using substantially Korman's method:

g_m = transconductance at operating point
g_{m0} = modulator transconductance at zero bias
E_g = modulator bias voltage at operating point
E_c = modulator cutoff bias voltage
e_r = modulator grid peak radio-frequency voltage
e_a = modulator grid peak audio-frequency voltage
e_0 = peak oscillator tank voltage
C_f = oscillator tank fixed capacitance
i_f = peak oscillator fixed capacitance tank current
i_m = peak modulator radio-frequency current
ω = angular frequency $= 2\pi f$
X_0 = reactance per leg of oscillator tank
w_m = band-width capability of modulator (cycles)
S_m = modulator sensitivity (cycles per volt)

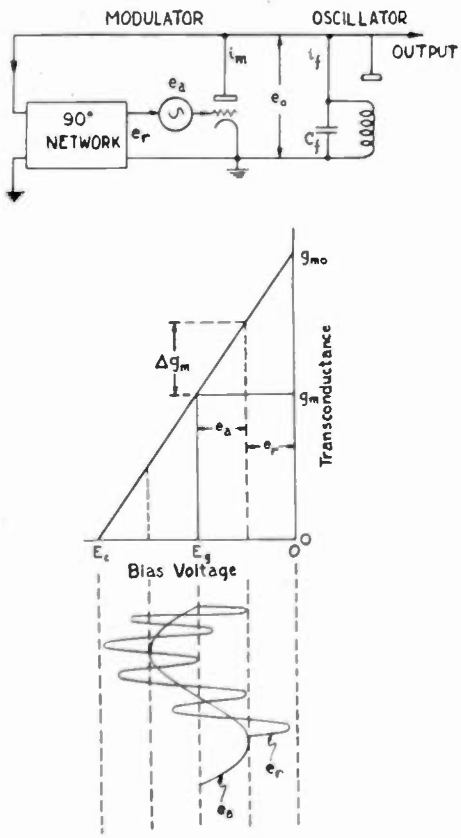


Fig. 1—Fundamental circuit of reactance-tube modulator.

Maximum modulator sensitivity occurs when the peak audio- and radio-frequency voltage swings on the modulator grids are equal, that is, when $e_a = e_r = \frac{1}{2}E_g$. Also, for the sake of simplicity, it will be assumed that the network always supplies the grids with voltage having a perfect 90-degree phase relationship during the audio cycle and that the in-phase component of modulator load on the oscillator may be neglected. The radio-frequency peak current to the modulator plate without modulation is

[1] RCA Manufacturing Company, Camden, N. J.

$$i_m = e_r g_m. \tag{1}$$

The oscillator peak radio-frequency current is

$$i_f = \frac{e_0}{X_0} = e_0 \omega C_f. \tag{2}$$

The fraction of tank current flowing through the modulators without modulation then is

$$\left(\frac{i_m}{i_f}\right)_0 = \frac{e_r g_m}{e_0 \omega C_f} \cdot \tag{3}$$

At full deviation (peak e_a), g_m becomes 50 per cent greater, or

$$\left(\frac{i_m}{i_f}\right)_{100} = \frac{3}{2}\,\frac{e_r g_m}{e_0 \omega C_f} \cdot \tag{4}$$

The fractional change of current from zero to full deviation is the difference between (4) and (3). The resultant fractional frequency change is one half of this for deviations of very small percentage of the oscillator frequency. Thus,

$$\frac{\Delta f}{f} = \frac{1}{2}\left[\left(\frac{i_m}{i_f}\right)_{100} - \left(\frac{i_m}{i_f}\right)_0\right]$$
$$= \frac{e_r g_m}{4 e_0 \omega C_f} = \frac{e_r g_m}{8\pi f e_0 C_f} \cdot \tag{5}$$

The maximum band width is then

$$w_m = 2\Delta f = \frac{e_r g_m}{4\pi e_0 C_f} = \frac{E_g g_m}{8\pi e_0 C_f} = k_m \frac{E_g g_m}{e_0 C_f} \tag{6}$$

in which k_m, the "modulator constant," is $1/8\pi$ or less, depending upon the percentage of the modulator g_m characteristic which is linear and usable for the *sum* of peak audio modulation voltage and direct-current bias shift with the most extreme automatic-frequency-control action. For an actual tube the truly linear portion of its transconductance characteristic may be but a fraction of this. The sensitivity is then

$$S_m = \frac{w_m}{E_g} \cdot \tag{7}$$

This derivation applies only to single-tube class A modulators and for the assumptions specified. For push-pull class A modulators the value given by (6) for w_m must be multiplied by two, and from (7) S_m also doubles. If the modulators are operated push-pull or "push-push" class B without grid current, a study of Fig. 1 will show that E_g and g_m may be replaced by E_0 and g_{m0} in (6), and similarly in (7) E_g may be replaced by E_c. The latter gives twice the sensitivity and four times the band-width capability of a single tube.

Modulated-Oscillator Drift

The term "drift" throughout this paper is used in a broad sense to mean the opposite of stability. It means instability due to any of the more common causes such

as temperature change, line voltage variation, humidity variation, and mechanical sensitiveness. The last two are considered purely problems of mechanical design, and may be minimized to almost any extent depending upon the elaborateness of design. Terms to take these factors into account, therefore, will not appear in the equations.

The inherent stability of push-pull reactance-modulated oscillators is found by experience to be limited primarily by irregularities and drift in the modulator circuits rather than by tube capacitance variation, etc., of the oscillator itself. This effect is taken into account by including S_m in the following equation for absolute (cycles, not percentage) drift:

$$\Delta F_0 = k_0 F_1 S_m \Delta t_0 \tag{8}$$

in which

ΔF_0 = drift of modulated oscillator without automatic frequency control.

F_1 = oscillator frequency

k_0 = oscillator drift constant

Δt_0 = range of oscillator tank temperature in operation

A factor to take into account line voltage variation is not included for push-pull modulators because their inherent balance makes them insensitive to all supply voltage variation;[2] that this is a fair assumption is well borne out in practice. The proper use of cathode-tap and screen-supply compensation on the oscillator itself to minimize line voltage effects is well covered in the literature.

It is commonly known that high-capcaitance tanks reduce drift in general, principally because the variable circuit and tube capacitances are a small proportion of the tank capacitance. This effect is included implicitly in the above equation in the term S_m which by (6) includes C_f in the denominator. Increasing C_0 will decrease the sensitivity of the modulator to grid-circuit disturbances, including intended modulation. If the modulator sensitivity is in any other way restored, for example by reducing oscillator tank voltage, the modulated oscillator again will be as vulnerable to grid-circuit disturbances as before. To realize the stabilizing effect of a high-C tank, it is essential, of course, to use minimum-drift coil construction.[3]

Substituting (7) for S_m, the inherent drift of the modulator-oscillator combination is

$$\Delta F_0 = k_0 \frac{w_m F_1}{E_g} \Delta t_0. \tag{9}$$

Note that the choice of single, push-pull class A or B modulators does not affect this expression as sensitivity is implicit in w_m. It is also important not to confuse w_m, the modulator band-width capability, with w_0, the modulated oscillator band width in use; w_0 may be equal to or less than w_m.

[2] M. G. Crosby, "Reactance tube frequency modulators," *RCA Rev.*, vol. 5, pp. 89–96; July, 1940.
[3] S. W. Seeley and E. I. Anderson, "UHF oscillator frequency stability consideration," *RCA Rev.*, vol. 5, pp. 77–88; July, 1940.

Modulated-Oscillator Asymmetry

Since the frequency varies as the square root of the capacitance in the tank, the inductance remaining constant, then for very small fractional changes of frequency $\Delta F/F$, the fractional change of capacitance must be $2\Delta F/F$. This is a very close approximation but for distortion consideration it must be more accurately evaluated. For explicitness a capacitive modulator will be assumed.

In Fig. 2, the modulator "synthetic" capacitance is broken into two components: one is fixed and considered to be part of the tank fixed capacitance C_f; the other, C_m, is varied by modulation and is just sufficiently large so that full positive C_m will produce deviation to "100 per cent modulation." When C_m is zero, the tank is on center frequency; when it is negative the swing is in the opposite direction. At the center frequency, resonance is expressed by

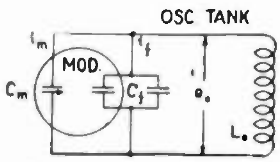


Fig. 2—Reactance-tube-modulator circuit for consideration of asymmetry.

$$\omega^2 = \frac{1}{L_0 C_f}. \tag{10}$$

The tank impedance relation is expressed by

$$C_f = \frac{i_f}{e_0 \omega} \tag{11}$$

and

$$\omega = \frac{e_0}{L_0 i_f}. \tag{12}$$

At a slightly higher frequency, resonance is expressed by

$$(\omega + \Delta\omega)^2 = \frac{1}{L_0(C_f - C_m)} \tag{13}$$

and C_f and C_m are in turn functions of instantaneous frequency. If i_{f1} is defined as the new current through C_f due to the change of frequency, i_f as the current before the frequency change, and i_m as the current added by the modulator capacitance C_m, then

$$C_f = \frac{i_{f1}}{e_0(\omega + \Delta\omega)} \quad \text{and} \quad -C_m = \frac{i_m}{e_0(\omega + \Delta\omega)}. \tag{14}$$

Substituting these values in (13) gives

$$(\omega + \Delta\omega)^2 = \frac{1}{\dfrac{L_0}{e_0}\left(\dfrac{i_{f1}}{\omega + \Delta\omega} + \dfrac{i_m}{\omega + \Delta\omega}\right)}. \tag{15}$$

If we may assume a constant tank voltage e_0, then

$$i_{f1} = \frac{\omega + \Delta\omega}{\omega} i_f \tag{16}$$

and
$$\frac{\omega + \Delta\omega}{\omega} = 1 + \frac{\Delta\omega}{\omega} = \frac{i_f}{i_f + \frac{\Delta\omega}{\omega} i_f + i_m} \tag{17}$$

$$\left(\frac{\Delta\omega}{\omega}\right)^2 i_f + \frac{\Delta\omega}{\omega}(2i_f + i_m) + i_m = 0.$$

Solution of the quadratic and examination for $i_m = 0$ indicate the choice of the positive square root of $(\Delta\omega/\omega)^2$ which is

$$\frac{\Delta\omega}{\omega} = \sqrt{\left(\frac{i_m}{2i_f}\right)^2 + 1} - \frac{i_m}{2i_f} - 1. \tag{18}$$

Assuming a constant tank voltage, constant phase applied to the modulator grids, and perfectly linear modulators, the fractional current deviations at peak frequency deviations on each side are equal, or

$$\left|\frac{-i_m}{2i_f}\right| = \left|\frac{+i_m}{2i_f}\right|. \tag{19}$$

We may calculate the fractional deviation at given instantaneous current deviations as shown in Table I.

TABLE I

$\frac{\Delta\omega}{\omega}$ for $-i_m$	$\frac{i_m}{2i_f}$	$\frac{\Delta\omega}{\omega}$ for $+i_m$
0.619	0.5	−0.381
0.220	0.2	−0.180
0.1050	0.1	−0.095
0.05125	0.05	−0.04875
0.0202	0.02	−0.0198
0.01005	0.01	−0.00995
0.0050125	0.005	−0.0049875
0.002002	0.002	−0.001998
0.0010005	0.001	−0.0009995
0.000500125	0.0005	−0.000499875
0.00020002	0.0002	−0.00019998
0.000100005	0.0001	−0.000099995
0.0	0.0	0.0

These values are plotted as linearity characteristics for five values of maximum fractional deviation on Fig. 3. As the fractional deviation is reduced, the expression approaches

$$\frac{\Delta\omega}{\omega} = \frac{i_m}{2i_f}$$

from which the fractional band width is related directly to modulator current swing and tank current by

$$\frac{w_0}{F_1} = 2\frac{\Delta\omega}{\omega} = \frac{i_m}{i_f}. \tag{20}$$

By using the distortion analysis of Chaffee[4] to calculate approximate distortion, from the curves of Fig. 3 (and from the above table when the distortion is too small to be shown on Fig. 3), this distortion may be listed as a function of fractional band width as shown in Table II.

[4] E. L. Chaffee, "A simplified harmonic analysis," *Rev. Sci. Instr.*, p. 384; October, 1936.

The somewhat startling result is apparent, that distortion is an approximately fixed percentage of fractional band width, i.e., 14 per cent. For example, a swing of plus and minus 100 kilocycles of a 50-megacycle modulated oscillator is a fractional band width of 0.4 per cent; therefore inherent distortion due to reactance curvature is $0.14 \times 0.4 = 0.06$ per cent. Obviously, if this swing is obtained by multiplying up from a lower

TABLE II

$\frac{W_0}{F_1}$	Distortion D	$\frac{DF_1}{w_0}$
Per Cent	Per Cent	
100	13.6	0.136
40	5.3	0.132
20	2.8	0.14
10	1.4	0.14
4	0.56	0.14
2	0.28	0.14
1	0.14	0.14
0.4	0.056	0.14
0.2	0.028	0.14
0.1	0.014	0.14

center frequency, the fractional band width and this component of distortion remain the same. Since the distortion calculated above is unnecessarily low at the present state of the art, it may be assumed that "beating up" from larger fractional band widths is permissible. If the circuit is so designed, the real limitation of band width of the modulated oscillator may be determined by

$$\frac{w_0}{F_1} = \frac{D}{B} \tag{21}$$

in which D = distortion permissible in modulated oscillator alone

$B = 0.14$ with a perfect modulator, or something larger for the practical modulator, allowing for nonlinearity, phase shift away from 90 degrees with modulation, etc.

MODULATED OSCILLATOR LINEARITY CHARACTERISTICS

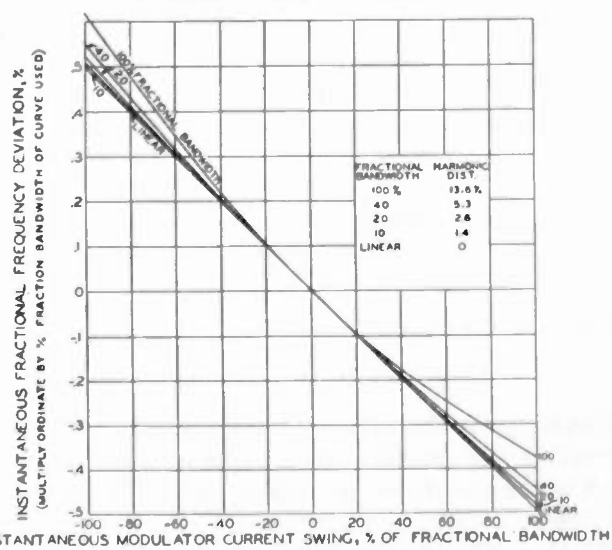


Fig. 3—Modulator linearity characteristics.

Consideration of Fig. 3 at first glance will indicate that the curves are far too good for the fractional band

widths listed. For example, the 100 per cent fractional-band-width curve refers to a swing of frequency, peak to peak, *equal* to the operating center frequency. If this band width were obtained by merely detuning the tank capacitor, the frequency would follow the familiar square-law relation, $F=1/2\pi\sqrt{LC}$, directly, resulting in much higher distortion. A reactance tube, however, does not behave like our familiar fixed inductive and capacitive reactances. Instead it produces a reactance in which the current is independent of any frequency variation across its terminals. Point-by-point consideration of the modulating cycle will reveal the compensating effect, a form of feedback, directly responsible for the lower inherent distortion. In this discussion the use of audio feedback is not taken into account, since the extremely good linearity of the modulators ordinarily does not require it. The output of the discriminator, properly filtered, is an integrated or direct-current value depending on the "center frequency" entering the discriminator.

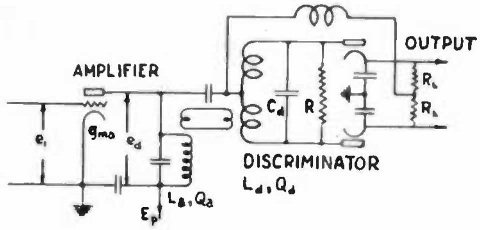


Fig. 4—Seeley discriminator circuit.

Discriminators

S. W. Seeley and D. E. Foster[5] have derived the sensitivity of the "Seeley"-type discriminator of Fig. 4 as

$$S_d = AL_dQ_d^2g_{ma}e_1 \text{ volts per cycle} \tag{22}$$

in which A = a factor evaluated in the article for various conditions of coupling and L_d/L_a ratios. (If we assume $L_d=L_a$, $Q_d=Q_a$, R_L is large compared to R, and the optimum coupling $k=0.86$, then $A=12$.)
L_d = discriminator secondary inductance
Q_d = discriminator secondary Q, including effect of load resistance R
g_{ma} = amplifier transconductance
e_1 = amplifier input root-mean-square voltage

This may, therefore, be rewritten, referring to Fig. 4, as

$$S_d = 12C_dR^2g_{ma}e_1 \text{ volts per cycle.} \tag{23}$$

Assuming for the sake of simplicity that with proper coupling the amplifier tube looks into an impedance k_1R where k_1 is the proportionality factor, then

$$S_d = \frac{12}{k_1}C_dR(k_1Rg_{ma}e_1) = \frac{12}{k_1}C_dRe_d \tag{24}$$

in which e_d is the discriminator input root-mean-square voltage. If we make the assumption that in transmitter design we shall be able to supply the necessary drive and can choose the necessary tube to make e_d swing to full supply voltage on peaks, then

$$S_d = \frac{12E_p}{\sqrt{2}\,k_1}C_dR. \tag{25}$$

The approximate band width[6] of the tuned circuit at 3 decibels down is

$$w_d = \frac{1}{2\pi RC_d} = \frac{2\pi L_d f^2}{R} = \frac{F_d}{Q_d}. \tag{26}$$

Equating RC_d in (25) and (26) gives

$$S_d = \frac{12E_p}{2\pi\sqrt{2}\,k_1w_d} = k_s\frac{E_p}{w_d} \tag{27}$$

in which k_s is a discriminator sensitivity constant taking into account the lesser sensitivity obtainable with better linearity (say 0.5 decibel down instead of 3). It is a figure of merit of the discriminator and preceding amplifier.

S_d is a maximum sensitivity and varies *inversely* as the band width because the amplifier is considered a constant voltage source. Note that line voltage affects sensitivity directly.

Discriminator drift may be assumed to be inversely proportional to tank capacitance for reasonable tank proportions, provided the capacitance added to increase stability has a zero temperature coefficient or is kept at constant temperature. For transmitter design we may consider the discriminator to be located in a small temperature-controlled oven.

If this were not the case, C_d in the following equation would appear, from self-controlled oscillator experience, as some root of C_d, probably the square or cube root.

$$\Delta F_d = \frac{k_dF_d\Delta t_d}{C_d} \text{ cycles} \tag{28}$$

in which ΔF_d = discriminator drift at the discriminator frequency
k_d = discriminator drift coefficient
F_d = discriminator frequency
Δt_d = range of discriminator tank temperature
C_d = discriminator tank capacitance

A factor for line voltage effects is not required for properly balanced discriminator circuits. The effect of line voltage variation is merely to change the sensitivity, not the center frequency. High C_d will always minimize drift if the coil construction is good, but for reasonable tank proportions over a wide frequency range C_d must vary approximately inversely with frequency, so that

[5] D. E. Foster and S. W. Seeley, "Automatic tuning, simplified circuits, and design practice," Proc. IRE. vol. 25, p. 295; March 1937.

[6] F. E. Spaulding, Jr., "Design of superheterodyne intermediate-frequency circuits," *RCA Rev.*, vol. 4, p. 490; April, 1940.

$$C_d = \frac{k_c}{F_d} \tag{29}$$

in which k_c is a "discriminator stiffness" coefficient. Then

$$\Delta F_d = \frac{k_d}{k_c} F_d^2 \Delta t_d. \tag{30}$$

The discriminator frequency-squared term indicates that *fractional* drift is proportional to frequency. If C_d in (28) had been introduced as $\sqrt{C_d}$ as mentioned above, the fractional drift would vary with the square root of frequency.

The Crosby[2] Automatic-Frequency-Control Circuit

In Fig. 5, if F_0 is the drift of the modulated oscillator without automatic frequency control, the equation of the feedback circuit, (whether $m_1F_1 = F_d$ as shown, or whether m_1F_1 beats with F_x to supply F_d to the discriminator), may be found by point-to-point drift consideration to be

$$\Delta F_0 - m_d m_1 S_d S_m \Delta F_1 = \Delta F_1$$

or

$$\Delta F_1 = \frac{\Delta F_0}{1 + S_d S_m m_1 m_d} \,. \tag{31}$$

This evaluation of drift relations within the feedback circuit, and the form of the fundamental drift equation as used in (40) have been previously used by J. L. Barnes[7] and N. I. Korman[1] in an early analysis of the Crosby transmitter circuit. For simplicity we may assume that the feedback factor $S_m S_d m_1 m_d$ is much greater than unity. If we consider the multiplier output to be the output of the circuit, m_1F_1, then the output drift simplifies to the approximation

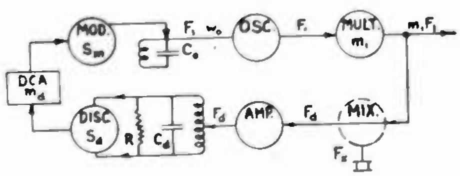


Fig. 5—Crosby automatic-frequency-control circuit.

$$\Delta(m_1F_1) \cong \frac{\Delta F_0}{S_m S_d m_d} \,. \tag{32}$$

On the other hand, if we write the output drift as a fraction of output frequency, substituting S_d from (27) and F_0/S_m from (8), the output drift due to modulated-oscillator drift alone is

$$\frac{\Delta(m_1F_1)}{m_1F_1} \cong \frac{k_0 w_d \Delta t_0}{k_s m_d m_1 E_p} \,. \tag{33}$$

The factor $m_1 m_d$ in the denominator shows the beneficial effect of using direct-current gain and of multiplying the oscillator frequency before feeding the discriminator. Band width of the discriminator is a function of the multiplication following it, and of the swing required at the antenna.

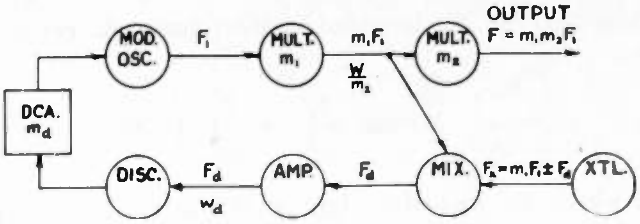


Fig. 6—Crosby transmitter circuit.

The drift due to the discriminator alone, by similar reasoning, may be written

$$\Delta F_1 = \frac{S_d S_m \Delta F_d}{S_d S_m m_1 + 1} \cong \frac{\Delta F_d}{m_1} \tag{34}$$

and the fractional drift, from (30), is

$$\frac{\Delta(m_1F_1)}{m_1F_1} \cong \frac{k_d}{k_c} \frac{F_d^2}{m_1F_1} \Delta t_d. \tag{35}$$

In the ratio F_d^2/m_1F_1 the reduction of drift by beating against a crystal is evident. The complete output drift from both sources is then

$$\frac{\Delta(m_1F_1)}{m_1F_1} \cong \frac{k_0}{k_s} \frac{w_d \Delta t_0}{m_1 m_d E_p} + \frac{k_d}{k_c} \frac{F_d^2 \Delta t_d}{F_1 m_1} \,. \tag{36}$$

The use of a direct-current amplifier of gain m_d in the feedback loop is obviously the same as an increase of m_1 so far as oscillator drift is concerned, and results in materially smaller drift.

The Crosby Transmitter Circuit

The output frequency drift ascribable to the oscillator alone is the same fractional drift as that at the output of the feedback loop and given by (33). We shall call this component D_0, substituting W/m_2 for w_d, to express the latter in terms of design requirements. Referring to Fig. 6, m_2 is the multiplication following the discriminator feed point. Then,

$$D_0 = \frac{k_0}{k_s} \frac{W}{m_1 m_2 m_d} \frac{\Delta t_0}{E_p} \,. \tag{37}$$

Expressing as D_d the fractional output drift due to the discriminator alone, we obtain from (35)

$$D_d = \frac{k_d}{k_c} \frac{F_d^2 \Delta t_d}{m_1F_1} \,. \tag{38}$$

By the same method used to obtain (31), the drift at m_1F_1 due to the crystal alone is

$$\Delta(m_1F_1) = \frac{m_d m_1 S_d S_m \Delta F_x}{m_d m_1 S_d S_m + 1} \cong \Delta F_x \tag{39}$$

and

$$\Delta F_x = k_x F_x \Delta t_x \tag{40}$$

[7] Formerly, RCA Manufacturing Company, Inc.

in which ΔF_x = actual drift in cycles
k_x = temperature coefficient of crystal
F_x = crystal frequency
Δt_x = range of temperature of crystal

From (39) the fractional output drift due to the crystal is

$$D_x \cong \frac{1}{m_1 F_1} k_x F_x \Delta t_x = \left(1 \pm \frac{F_d}{m_1 F_1}\right) k_x \Delta t_x \tag{41}$$

in which the negative sign applies when $F_x < F/m_2$ and vice versa. The output drift in cycles is then $F(D_0 + D_d + D_x)$ or

$$\text{total drift} = \frac{F_1 W}{m_d} \frac{k_0 \Delta t_0}{k_s E_p} + \frac{m_2 F_d^2 k_d}{k_c} \Delta t_d + (F \pm m_2 F_d) k_x \Delta t_x. \tag{42}$$

As m_2 is increased, F/m_2 decreases to a point beyond which the crystal must be operated on the high side of F/m_2 to obtain the required F_d. If we minimize the last two terms with respect to F_d, we obtain

$$\text{optimum } F_d = \frac{k_c k_x \Delta t_x}{2 k_d \Delta t_d}. \tag{43}$$

This expression gives the optimum discriminator frequency when the crystal is operated on the low side of F/m_2. If operated on the high side, minimum F_d always produces minimum drift. This optimum F_d occurs because the crystal frequency, and therefore its drift, is a function of F_d and m_2. An increase of F_d will reduce drift caused by the crystal but this advantage is later overshadowed by the F_d^2 term of the discriminator itself, resulting in increased drift.

For the case where the crystal is operated on the low side of F/m_2, F_d from (43) may be put into (42) to give

$$\text{total drift}_L = \frac{F_1 W}{m_d} \frac{k_0 \Delta t_0}{k_s E_p} + F k_x \Delta t_x - \frac{m_2 k_c (k_x \Delta t_x)^2}{4 k_d \Delta t_d}. \tag{44}$$

Because of the situation just pointed out, this equation applies only when

$$\frac{F}{m_2} - F_d > 0 \quad \text{or} \quad m_2 < \frac{F}{F_d} \tag{45}$$

in which F_d may be evaluated by (43).

At this point it is necessary to put numbers into the equations in order to draw conclusions from the result. The following tentative values have been obtained from experimental measurements on an early model Crosby exciter. Some constants obviously should be as high and some as low as possible (for example, the crystal drift coefficient should be low), some are fixed by requirements of the Federal Communications Commission, and others may be chosen economically at the designer's discretion.

$$\left.\begin{array}{ll}
k_x = 10^{-6} & \text{(one cycle/Mc./°C crystal)} \\
k_x \Delta t_x = 5 \times 10^{-6} & \text{(above in 0.5°C oven)} \\
k_d = 1.5 \times 10^{-15} & \text{(20 cycles/Mc./°C; 75 } \mu\mu\text{f)} \\
k_c = 0.001 & \text{(1000 } \mu\mu\text{f at 1000 kc)} \\
\Delta t_d = 0.5°\text{C} & \text{(discriminator in 0.5°C oven)} \\
k_0 = 16 \times 10^{-9} & \text{(40 cycles/Mc/°C at 4.5 Mc.} \\
 & \quad w_0 = 60 \text{ kc, } E_g = 24\text{v)} \\
k_s = 0.56 & \text{(0.53 v/kc, } e_a = 40 \text{ v r-m-s, } w_d = 60 \text{ kc)} \\
F = 50 \times 10^6 & \text{(top of 42- to 50-Mc. broadcast band)} \\
W = 0.2 \times 10^6 & \text{(plus and minus 100-kc swing)} \\
\Delta t_0 = 40°\text{C} & \text{(maximum ambient temperature range} \\
 & \quad \text{of 5°C to 45°C)} \\
E_p = 250 \text{ v} & \text{(for receiving tubes)}
\end{array}\right\} \tag{46}$$

Using the above figures in (43) we obtain

$$\text{optimum } F_d = \frac{0.001 \times 5 \times 10^{-6}}{2 \times 1.5 \times 10^{-15} \times 0.5} = 3.33 \text{ Mc.} \tag{47}$$

Maximum m_2 from (45) is

$$m_2 = \frac{50}{3.33} = 15. \tag{48}$$

Then from (46) and (44)

$$\text{total drift}_L = \frac{45750}{m_d m_1 m_2} + 250 - 8.33 m_2 \text{ cycles.} \tag{49}$$

These results are shown by the solid lines of Fig. 7.

From (42) and (46), if the crystal is operated on the high side of F/m_2,

$$\text{total drift}_H = \frac{45750}{m_d m_1 m_2} + 250 + m_2(\tfrac{3}{4} F_d^2 + 5 F_d). \tag{50}$$

From this, the dotted curves of Fig. 7 are plotted. Note that over-all drift is the sum of two drift components: one a function of F_d and the other a function of $m_d m_1$. Both are functions of m_2. The curves show the obvious reduction of drift by increasing the multiplication following as well as preceding the discriminator.

Fig. 7, then, indicates the proper choice of multiplying factors at various locations in the circuit, when the constants of (46) are used. The constants, in turn, are obtained from the band width and output frequency requirements of broadcast service, from practically obtainable modulator and discriminator sensitivities, from practically obtainable tank stability, and for the assumption that only the discriminator and crystal will be located in temperature-controlled ovens.

Equation (42) was checked against the experimental conditions existing in a model of the RCA MI-19407 frequency-modulated exciter,[8] in which a somewhat different set of constants from those of (46) was used in equation (42), and the over-all drift calculated to be 835 cycles. This equipment, when subjected to a 40-degree centigrade heat cycle, a 30 to 95 per cent humidity cycle, and a plus-or-minus 15 per cent line voltage cycle simultaneously for worst conditions, showed a maximum drift range of 900 cycles. This may be considered as plus or minus 450 cycles or 4.4 times as good as the plus or minus 2000 cycles required by the Federal Communications Commission.

[8] E. S. Winlund, "FM engineering considerations, part 2," *FM Magazine*, vol. 1, no. 10; August, 1941.

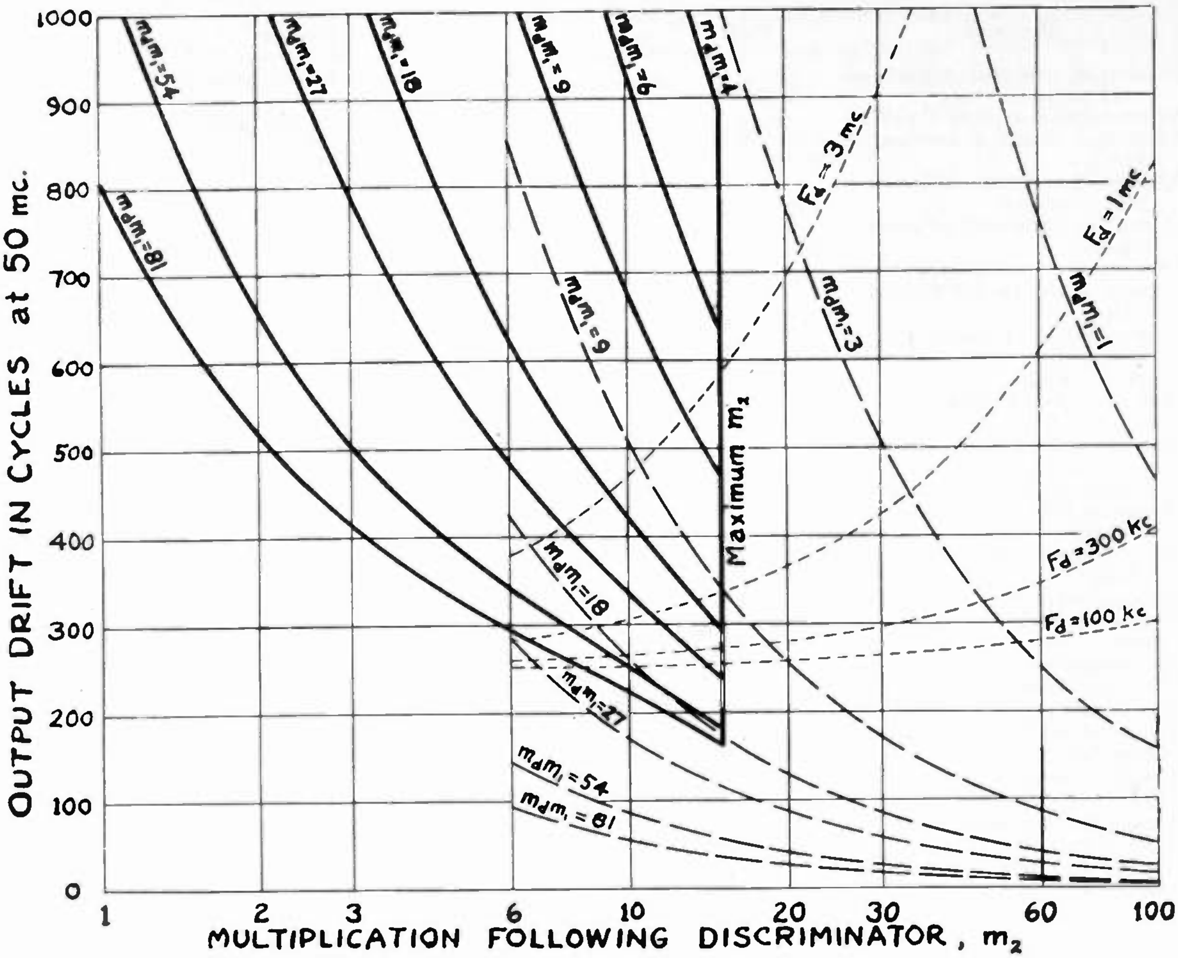


Fig. 7—Crosby transmitter circuit drift curves.

When crystal is on low side of modulated signal, use solid curves for total drift. Optimum discriminator frequency assumed to be 3.33 megacycles.
When crystal is on high side of modulated signal, use dotted curves for desired F_d and m_dm_1. Total drift is the sum of the two.
m_dm_1 = gain in feedback loop
F_d = discriminator frequency.

Conclusions

By proper attention to design and choice of frequency multiplication in individual portions of the circuit, applications of the Crosby automatic-frequency-control circuit are capable of providing stability of the same order of magnitude as a crystal.

The Crosby transmitter circuit is capable of high stability with a relatively high modulated oscillator frequency, provided a simple heat oven or a high-gain direct-current amplifier is used in conjunction with the proper circuits.

Factors which aid stability are 1. low initial frequency, 2. direct-current gain in feedback loop, 3. high discriminator power, 4. low-drift crystals, 5. low-drift tank construction, and 6. heat-oven regulation of proper circuits. The use of but two or three of these expedients, the choice being one of design simplicity, will usually bring the drift within requirements.

It is recognized that many assumptions have been made in this analysis in order to arrive at a concrete and practically usable result. While several of these may be very approximate over a wide frequency range, the treatment provides a definite foothold for comparing and designing circuits of this type.

Acknowledgment

The assistance and suggestions received from my coworkers during this analysis and during the design of the commercial equipment is very gratefully acknowledged.

Appendix

For convenience in following the analysis, a list of all symbols follows.

Symbol	Definition	Units	First used with equation
A	Seeley coupling factor	factor	23
B	Asymmetry factor	fraction	21
C_d	Discriminator tank capacitance	farads	23
C_f	Modulated-oscillator tank fixed capacitance	farads	2, 10
C_m	Modulated-oscillator tank modulated capacitance	farads	13

Symbol	Definition	Units	First used with equation
D	Distortion, root mean square, referred to fundamental	fraction	21
DCA	Direct-current amplifier of gain m_d	—	34
D_d	Discriminator drift at antenna frequency	cycles	41
D_0	Modulated-oscillator drift at antenna frequency	cycles	40
D_x	Crystal-oscillator drift at antenna frequency	cycles	44
e_1	Discriminator-amplifier input root-mean-square radio-frequency voltage	volts	22
e_a	Peak modulator grid audio voltage	volts	1
e_d	Discriminator input root-mean-square voltage	volts	24
e_0	Peak modulated-oscillator tank voltage	volts	2, 11
e_r	Peak modulator grid radio-frequency voltage	volts	1
E_c	Modulator cutoff bias voltage	volts	7
E_g	Modulator bias voltage at operating point	volts	6
E_p	Discriminator-amplifier supply voltage	volts	25
F	Antenna center frequency	cycles	45
F_1	Modulated-oscillator center frequency	cycles	8
F_d	Discriminator center frequency	cycles	26, 31
F_x	Crystal-oscillator frequency	cycles	43
ΔF_d	Discriminator drift at frequency F_d	cycles	31
ΔF_0	Modulated-oscillator drift at frequency F_0	cycles	8
ΔF_x	Crystal-oscillator drift at frequency F_x	cycles	43
g_m	Modulator transconductance at operating point	mhos	1
g_{ma}	Discriminator amplifier transconductance	mhos	22
g_{m0}	Modulator transconductance at zero bias	mhos	7
i_f	Peak modulated-oscillator fixed-capacitance tank current	amperes	2, 12
i_{f1}	Modulated-oscillator i_f at peak deviation	amperes	14
i_m	Peak modulator radio-frequency current	amperes	14
k_1	Discriminator impedance transfer constant	factor	24
k_c	Discriminator stiffness coefficient	farad-cycles	32
k_d	Discriminator drift coefficient	farads/°C	31
k_m	Modulator sensitivity coefficient	fraction	6
k_0	Modulated-oscillator drift coefficient	volts/cycle/°C	8
k_s	Discriminator sensitivity coefficient	factor	30
k_x	Crystal-oscillator drift coefficient	1/°C	43
L_a	Discriminator-amplifier inductance	henrys	22
L_d	Discriminator tank inductance	henrys	22
L_c	Modulated-oscillator tank inductance	henrys	10
m_1	Frequency multiplication from modulated-oscillator to discriminator	integer	34
m_2	Frequency multiplication following discriminator	integer	40
m_d	Direct-current amplifier gain in feedback loop	ratio	34
Q_a	Discriminator-amplifier tank Q	ratio	22
Q_d	Discriminator tank Q	ratio	22
R	Discriminator tank load resistance, radio-frequency	ohms	23
R_L	Discriminator direct-current load resistance	ohms	22
S_d	Discriminator sensitivity	volts/cycle	22
S_m	Modulator sensitivity	cycles/volt	7
Δt_d	Discriminator temperature range	°C	31
Δt_0	Modulated-oscillator temperature range	°C	8
Δt_x	Crystal-oscillator temperature range	°C	43
w_d	Discriminator band width at frequency F_d	cycles	28
w_m	Modulator band-width capability at frequency F_1	cycles	6
w_0	Modulated-oscillator band width at frequency F_1	cycles	20
W	Band width at antenna frequency F	cycles	40
ω	Modulated-oscillator angular frequency $2\pi F_1$	radians	2
$\Delta\omega$	Modulated-oscillator angular deviation $2\pi\Delta F_1$	radians	13
X_0	Reactance per leg of modulated-oscillator tank	oh s	2

Observations of Frequency-Modulation Propagation on 26 Megacycles

MURRAY G. CROSBY†, MEMBER, I.R.E.

Summary—*A simplified analysis is given which shows that the distortion occurring with two-path transmission of frequency modulation consists of the introduction of a new modulation frequency which is frequency-modulated and is not harmonically related to the fundamental modulation frequency. Results of observations at Riverhead, Long Island, New York, on 26.3-megacycle frequency-modulated transmissions from W9XA[1] at Kansas City are shown and discussed.*

IN THE observations to be described here, conditions were rather ideal for study of the type of distortion produced by multipath effects on frequency modulation. As would be expected for such a high radiation frequency as 26.3 megacycles, and the short distance of 1150 miles, transmission was such that there were comparatively few paths with a relatively small time delay between them. During the major portion of the observations, transmission appeared to be either over a single path or what appeared to be two paths with an extremely short time delay between them. The rest of the time, transmission was obviously over two paths with an appreciable time delay between them. Hence the simple conditions of only a single path or two paths were available for study without the complications introduced by the many paths encountered

* Decimal classification: R148.1×R414. Original manuscript received by the Institute, April 24, 1941. Presented, Summer Convention, Detroit, Mich., June 25, 1941.

† R.C.A. Communications, Inc., Riverhead, L. I., N. Y.

[1] Owned and operated by the Commercial Radio Equipment Company, 7134–36 Main Street, Kansas City, Missouri.

at the lower radiation frequencies which were used in previous tests.[2]

Before proceeding with the discussion of the results of the observations, it will be well to consider the type of distortion which this type of transmission may produce. A mathematical analysis of the manner in which the distortion is formed has been given before.[2] The following is a simplified study intended to give a physical picture of the type of distortion which takes place.

Simplified Analysis

In the multipath transmission to be considered here, it will be assumed that there are two paths with an amplitude ratio R and a time delay between the two paths of D seconds.

The upper part of Fig. 1 is a graph of the instantaneous frequencies of a sinusoidally frequency-modulated wave transmitted over two paths. The ordinates of the graph are plotted in frequency deviation either side of the carrier frequency F_c. Wave B arrives D seconds later than wave A so that the phase delay α between the two modulating waves is equal to $2\pi DF_m$, where F_m is the frequency of the modulating wave.

It will be noted that the effect of the time delay between the two waves is to cause the frequency variations to be out of synchronism so that a beat note will be formed. This beat note will be equal to the difference between the instantaneous frequencies of the two waves. Thus at the time b in Fig. 1, the beat note will be equal to the frequency difference $y-x$. At the time c, it will be equal to $z-w$ and at the time d the difference is zero. The frequency difference changes sign, or reverses phase, at the zero points a, d, e, etc.

It can be seen that the presence of two waves of different frequency produces a resultant wave which is amplitude- and phase-modulated. Thus the effect of the multipath transmission is to superimpose an amplitude modulation on the wave amplitude and a phase modulation on the phase or frequency of the wave. In the case to be examined here, it will be assumed that the amplitude ratio R of the two waves is less than unity so that the superimposed amplitude modulation will be less than 100 per cent and may be removed by the limiter in the frequency-modulation receiver. The phase-modulation component is therefore the only one which need be considered.

The phase-modulation component of the resultant is to be received on a frequency-modulation receiver so that, in order to determine the output of the frequency-modulation receiver when it is fed by such phase-modulated distortion, the effective frequency deviation of the phase modulation must be determined. In order to do this, the beat-note frequency and the degree of phase modulation must be evaluated. From these two values, the effective frequency deviation may be ascertained. The beat-note frequency is equivalent to the modulating frequency of the phase-modulation component and is equal to the difference between the instantaneous frequencies of the two asynchronous waves. This difference is merely the difference between the amplitude of the two sine waves of frequency deviation:

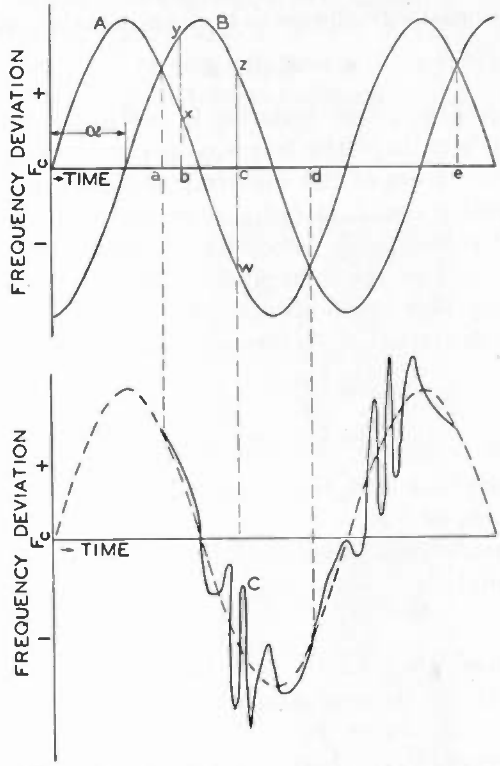


Fig. 1—Manner in which the frequency variations of wave A and time-delayed wave B combine to produce a resultant wave C having a frequency-modulated distortion component.

$$f_A - f_B = F_d \sin pt - F_d \sin (pt + \alpha) \tag{1}$$

or,

$$F_{md} = 2F_d \sin \frac{\alpha}{2}\left[\cos\left(pt + \frac{\alpha}{2}\right)\right] \tag{2}$$

where f_A and f_B are the respective instantaneous frequencies, F_d is the peak frequency deviation of the original frequency-modulated wave $p = 2\pi F_m$, and F_{md} is the modulation frequency of the distortion component.

From (2), it can be seen that the modulation frequency of the distortion component is sinusoidally varied between the limits of plus and minus $2F_d \sin \alpha/2$. It is thus a frequency-modulated distortion which sweeps through zero cycles and has a maximum frequency dependent upon the applied frequency deviation F_d and the phase delay between the modulation frequencies of the two paths α. Since $\alpha = 2\pi DF_m$, the maximum frequency is in turn dependent upon the time delay between paths and upon the original applied modulation frequency. The maximum frequency goes through maxima and minima in accordance with $\sin \alpha/2$ as α is varied by a variation of either D or F_m.

[2] Murray G. Crosby, "Frequency modulation propagation characteristics," Proc. I.R.E., vol. 24, pp. 898–913; June, 1936.

The maximum possible beat note is equal to twice the applied frequency deviation and occurs when α is 180, 540, etc., degrees.

The magnitude to the phase deviation of the resultant vector composed of two waves having a beat note can easily be shown to be approximately given by

$$\phi = R \text{ (for } R < 1) \qquad (3)$$

where ϕ is the phase deviation in radians, and R is the amplitude ratio of the two waves. Hence the effective phase deviation of the superimposed distortion is approximately equal, in radians, to the amplitude ratio of the two paths. In the case to be considered here, R is close to, but less than, unity.

It may also be shown that the effective peak frequency deviation of a phase-modulated wave is given by

$$F_d = \phi F_m \qquad (4)$$

where F_d is the effective peak frequency deviation, ϕ is the peak phase deviation of the phase-modulated wave in radians, and F_m is the modulation frequency of the phase-modulated wave. Substituting (3) and (4) gives the formula

$$F_d = RF_m \qquad (5)$$

which gives the effective peak frequency deviation F_d, produced by the resultant of two sinusoidal waves having an amplitude ratio R and a difference, or beat-note frequency, F_m. Inserting the beat-note frequency of (2) for F_m in this formula gives

$$F_{dd} = 2RF_d \sin\frac{\alpha}{2}\left[\cos\left(pt + \frac{\alpha}{2}\right)\right] \qquad (6)$$

where F_{dd} is the peak frequency deviation of the distortion component.

Equation (2) gives the relation for the modulation frequency of the distortion component and (6) gives its peak frequency deviation. The resulting relation for the frequency of the received carrier therefore has the form

$$f = F_c + F_d \sin pt - F_{dd} \sin (2\pi F_{md} t) \qquad (7)$$

in which F_c is the carrier frequency and F_{dd} and F_{md} are given by (6) and (2), respectively. The output of the frequency-modulation receiver which is fed by this frequency-modulated wave will be

$$J = kF_d\left[\sin pt - \left\{2R \sin\frac{\alpha}{2} \cos\left(pt + \frac{\alpha}{2}\right)\right\} \cdot \sin 2\pi\left\{2F_d \sin\frac{\alpha}{2} \cos\left(pt + \frac{\alpha}{2}\right)\right\} t\right]. \qquad (8)$$

It can thus be seen that the multipath distortion appears as a superimposed modulation which is sinusoidally frequency-modulated and has a depth of modulation which is proportional to this variable modulation frequency. Hence, as the distortion is modulated through zero frequency, the amplitude passes through zero, and as it is modulated to its highest frequency, the amplitude increases to its maximum value. The maximum value of the distortion frequency deviation will be slightly less than twice the original frequency deviation. This value will occur when R is close to unity and α is 180 degrees, 540 degrees, etc. Hence, the amplitude of the distortion component may rise to as high as approximately twice the amplitude of the original modulation frequency.

Wave C of the lower graph of Fig. 1 shows how the distortion modulation is superimposed on the original modulated wave which is shown dotted. The maximum frequency of the distortion modulation occurs at the time c and has a value equal to $2F_d \sin \alpha/2$. At times on either side of this maximum frequency point, the frequency and the amplitude of the distortion component decreases towards zero at the times a and d.

Since the superimposed distortion is in the form of a phase modulation like the combination of a carrier with noise, the frequency-modulation system discriminates somewhat against the multipath effects just as it does against noise. This discrimination shows up as a reduction of distortion as the modulation frequency is lowered below the frequency which causes α to be 90 degrees or the quantity πDF_m to be equal to $\pi/2$. However, such an effect is dependent upon the desired signal being stronger than the undesired. Hence for the discrimination against multipath effects to be effective, the signal arriving over one path must be at least twice as strong as the resultant of the others. For the case of ionospheric transmission, this can only happen occasionally. Furthermore, with the values of D normally encountered in ionospheric transmission (from a few microseconds to several milliseconds), the maximum allowable F_m would have to be quite small in order for the quantity πDF_m to be less than $\pi/2$.

In the wave form above discussed, it was assumed that the amplitude ratio of the two paths was less than unity. Of course there is nothing to insure this condition in actual practice so that the condition of 100 per cent amplitude modulation, which occurs when the ratio is unity, would frequently occur. When 100 per cent amplitude modulation occurs, the limiter brings up the noise during the signal minima. As will be shown in the results of the observations, the signal minima occur at the peak of the distortion-frequency cycles. The noise is therefore introduced during these peaks.

Wave-Form Observations on W9XA

The observations were made during the months of June and July, 1940, during which time W9XA radiated frequency-modulation transmissions of tone and program modulation on 26.3 megacycles. Although this frequency is too high for consistent transmission over the 1150-mile path between Kansas City and the receiving location at Riverhead, L. I., N. Y., there were many days of good transmission. Signal strengths

up to about 15 microvolts per meter were received.

Modulation tones of 100, 400, and 2000 cycles were applied with a frequency deviation of about 26 kilocycles. The wave form was observed on an oscilloscope which was photographed with a 16-millimeter motion-picture camera.

The oscillograms of Fig. 2 show the typical types of wave-form distortion obtained on the various tone frequencies. Aside from the periods during which there appeared to be single-path transmission with little or no distortion, there were two separate conditions of distortion which were observed. The first was a condition of an apparent two-path transmission with an extremely short time delay of a few microseconds between paths. The second was a condition of an apparent two-path transmission with a time delay of about 100 microseconds between paths. The condition of single-path transmission predominated during about one half of the days signals were received. The two conditions of distortion each predominated during about one fourth of the observations.

The result of the first condition in which the time delay was a few microseconds was the introduction of a single dip or jag in each half cycle of the wave. Most of the time the signal dropped below the noise level during this dip so that it appeared as though there were a "hole" in the ether in which there was zero transmission. Typical instances of this condition are shown in the oscillograms of Fig. 2B, C, D, K, and L. In the case of Fig. 2B, there is also a slight indication of the existence of a path of longer time delay as shown by the higher frequency distortion superimposed between the two noise peaks. In Fig. 2C, the peak of the wave passed through the "hole" and let the noise through, while in Fig. 2D only the peak of the wave came through. In Fig. 2K, the amplitude ratio of the two paths was not sufficiently close to unity to cause the resultant signal to cancel to a value below the noise level. Fig. 2L is similar to Fig. 2B except that the tone frequency was 2000 cycles instead of 400.

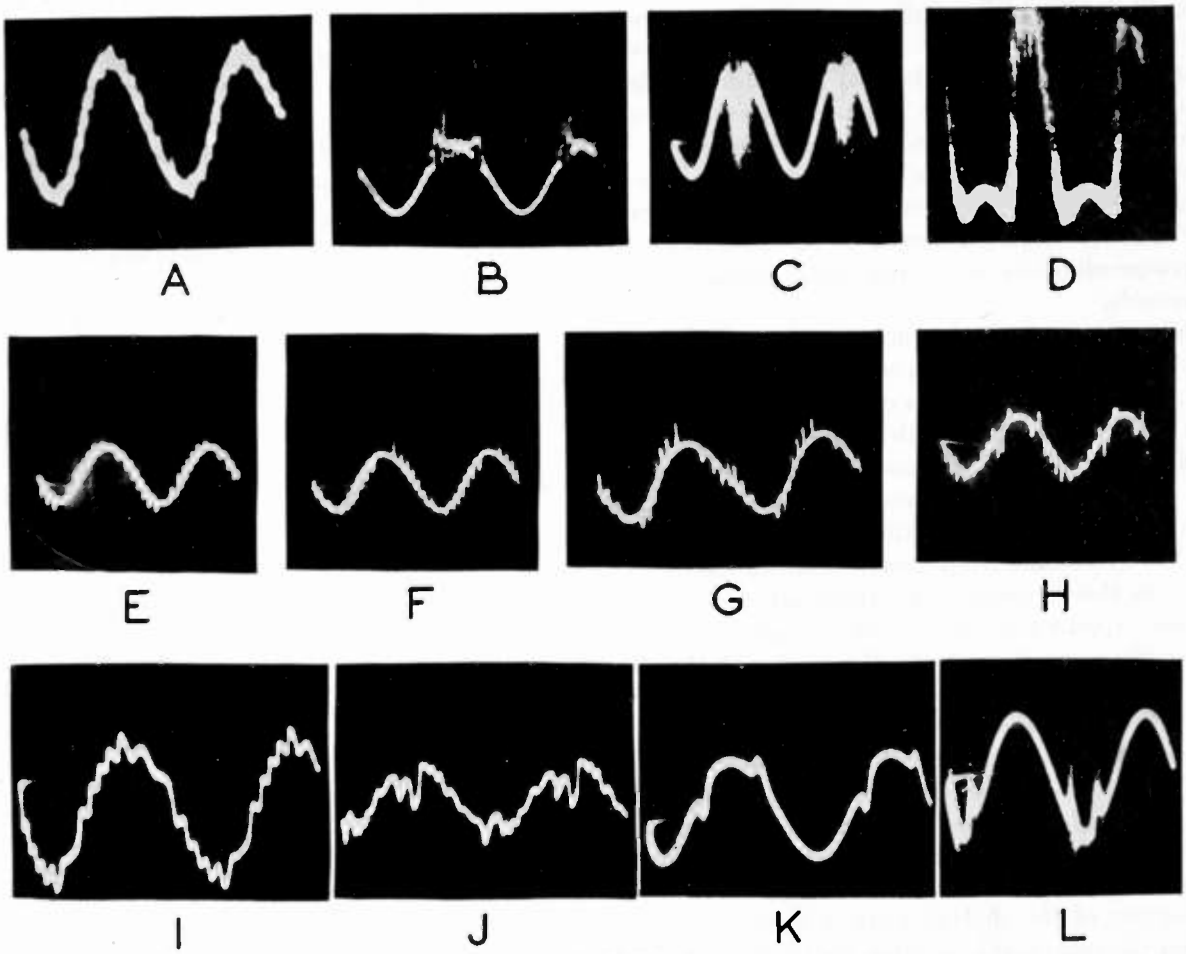


Fig. 2—Oscillograms of the wave forms received from the W9XA frequency-modulation transmissions. The modulation frequencies used for these oscillograms are: A—100 cycles; B, C, D, E, F, G, and H—400 cycles; I, J, K, and L—2000 cycles.

When this first condition existed, the superimposed distortion beat note between the two waves separated by a time delay apparently rose to something between one-half cycle and a full cycle. The sharpness of the peaks produced might indicate a beat frequency of higher than one half to one cycle. However, this sharpness is obviously due to the peculiar peakedness of the wave form obtained from the frequency-modulation

component of the beat note when the amplitude ratio is near unity.[3]

Since, during this first condition, the dip also occurs as a "hole," it is apparent that the condition of out-of-phase radio-frequency combination of the two waves produces the dip. Thus, the location of the dip with respect to the original modulation cycle is dependent upon the particular radio-frequency phase relation existing at the time. As this radio-frequency phase relation varies, the location of the dip or "hole" shifts from one point on the cycle to another. This shifting was usually taking place continuously in the observations.

The time delay between the two paths for this first condition was evaulated by equating the maximum superimposed distortion modulation frequency to $2F_d \sin \pi DF_m$ and solving for D. Assuming F_{md} as one cycle gave a time delay of 3 microseconds for the case of Fig. 2K. If F_{md} were assumed to be one-half cycle, the corresponding value of time delay would be 1.5 microseconds.

In the second condition, the time delay was sufficiently large so that an appreciable beat note was formed. As can be seen from a comparison of Figs. 2A, E, F, G, H, and I, the depth of modulation of this superimposed distortion increased as the original modulation frequency was increased. The oscillogram of Fig. 2A shows about the maximum amount observed on the 100-cycle tone frequency. When the distortion occurred on this frequency, the amplitude ratio of the paths was apparently very close to unity because noise usually came through on the peaks of the distortion modulation.

Figs. 2E and F show adjacent frames of 16-millimeter motion pictures taken at 16 frames per second so that they are 1/16th of a second apart in time. It is apparent from the reversal of the phase of the distortion modulation, that the amplitude ratio of the two paths must have gone through unity between frames. The existence of the sharply peaked wave form indicates that the ratio was near unity for both frames and the reversal of the direction of the sharp peak shows that the stronger path of one frame became the weaker path of the other.[3] Hence, the relative amplitudes of the two paths must have been varying quite rapidly.

Figs. 2G and H show an indication of change in relative path amplitudes during a cycle. It will be noted that the direction of the sharp peaks of the distortion wave form reverses near the peak of the cycle. It is apparent from this that the two paths had different frequency characteristics so that at one frequency one path was predominant and at another the other path was predominant.

Fig. 2I shows the distortion that was quite typical of the 2000-cycle tone during this second condition of time delay. The maximum modulation frequency of the distortion F_{md} was about 32,000 cycles for this instance. This modulation frequency is higher than the receiver was designed to pass. However, since there was only one high-quality audio transformer in the audio system, this frequency was passed although at reduced amplitude. Equating this value of F_{md} to $2F_d \sin \pi DF_m$ gives a time delay of 105 microseconds for this instance. Applying the same procedure to the 400-cycle case of Fig. 2F gives a value of F_{md} of 7000 cycles and a time delay of 106 microseconds.

Fig. 2J shows a case in which there were apparently more than two paths. This condition was quite rare.

Program Observations

During the major portion of the program observations, transmission was apparently over a single path at which time the only effect noticed was the variability of the signal-to-noise ratio as the signal faded.

For the first condition of distortion, which produced the single dip in the tone wave form, there were a few observations of program modulation. These observations were characterized by the introduction of spurts of noise as the modulation was applied. Sometimes these spurts were of short time duration so that the annoyance was slight, while at other times the band width of the "hole" was apparently wide enough to remove a major portion of the modulation cycle and the result was severe distortion and noise.

It happened that during the short periods of the second condition of longer time delay between paths, the transmissions consisted of tone modulation so that no program observations for this condition were obtained.

Static Reception

On several occasions there was enough static to be heard on 26 megacycles. The static consisted of "bats" and "crashes" which came through with the usual sound on the amplitude-modulation receiver, but sounded much different on the frequency-modulation receiver. The effect of a crash was to eliminate the signal for the duration of the crash and substitute a steady hiss like tube hiss. The sound was the same as a momentary loss of carrier and could be duplicated by momentarily shorting the antenna. Usually the effect was inappreciable unless the strength of the crash was great enough to blot out the signal momentarily. On one occasion, the receiver was tuned to W1XOJ on 43.0 megacycles and the same static was observed.

It is apparent that this sound of the static is due to the noise-silencing properties of the receiver for the

[3] Murray G. Crosby, "Frequency modulation noise characteristics," Proc. I.R.E., vol. 25, pp. 472–514; April, 1937. Fig. 4 on page 486 shows the calculated wave form of the frequency variation produced by the combination of two sinusoidal waves of various amplitude ratios. When the amplitude ratio is near unity, the wave form is sharply peaked on one side and flattened on the other. Interchanging the relative amplitudes of the two waves reverses the polarity of the resulting wave form. Interchanging the relative frequencies of the two waves has the same effect. The result of an interchange of the relative amplitudes or frequencies of the two waves is thus to cause upward peaks to become downward and vice versa.

condition of noise with a greater peak voltage than that of the carrier. Under this condition, the noise discriminates against the signal so that the frequency-modulation improvement is working against the signal instead of against the noise. Also, due to the "frequency-limiting" property of the frequency-modulation receiver, the irregularities of the static impulse are limited off so that the sound is smooth like that of tube hiss.

Conclusions

The simplified analysis of the two-path case shows that the distortion consists of the introduction of a new modulation frequency which is frequency-modulated from zero to a frequency as high as twice the applied frequency deviation. The amplitude of the distortion component is proportional to the modulation frequency of the distortion component and may rise to a resulting frequency deviation as high as approximately twice that of the original modulation frequency

There were three conditions of transmission which apparently existed during the observations on W9XA. The first was a single-path transmission in which the only effect noticeable was the variation in signal strength. This condition existed on about one half of the days signals were received. The second was a two-path transmission with a short time delay of a few microseconds between paths. This second condition produced a single dip in the cycle of the tone modulation and caused the introduction of spurts of noise with modulation on program modulation. It prevailed on about one fourth of the days signals were received. The third condition was a two-path transmission with a time delay of about 100 microseconds between paths. This condition caused the introduction of frequency-modulated distortion components up to 32,000 cycles when the modulation tone was 2000 cycles. It took place on one fourth of the days signals were observed.

Acknowledgment

The author gratefully appreciates the kind co-operation of Mr. Everett L. Dillard, manager of the Commercial Radio Equipment Company at Kansas City, Missouri, in arranging special transmissions from W9XA.

The assistance of the author's associates in RCA Laboratories at Riverhead, Long Island, New York, is also gratefully acknowledged.

High-Frequency Radio Transmission Conditions June, 1941, with Predictions for September, 1941*

NATIONAL BUREAU OF STANDARDS, WASHINGTON, D.C.

The radio transmission data herein are based on observations at Washington, D. C., of long-distance reception and of the ionosphere. Fig. 1 gives the June average values of maximum usable frequencies, for undisturbed days, for radio transmission by way of the regular layers of the ionosphere. The maximum usable frequencies were determined by the F layer at night and by the E, F_1, and F_2 layers during the day. Fig. 2 gives the expected values of the maximum usable frequencies for radio transmission by way of the regular layers, average for undisturbed days, for September, 1941. Average critical frequencies and virtual heights of the ionospheric layers as observed at Washington, D.C., during June are given in Fig. 3.

TABLE I

Ionospheric Storms (approximately in order of severity)

Day and hour E.S.T.	h_F before sunrise (km)	Minimum $f_F{}^0$ before sunrise (Mc)	Noon $f_{F_2}{}^0$ (Mc)	Magnetic character[1]		Ionospheric character[2]
				00–12 G.M.T.	12–24 G.M.T.	
June						
12 (from 0200)	322	2.0	5.0	2.1	2.2	4
13	342	2.0	<4.5	3.8	3.4	6
14	308	1.8	4.9	2.9	2.1	3
15 (through 0900)	340	<1.6	—	3.5	3.0	4
17 (from 1600)	—	—	—	2.0	3.2	5
18	312	2.8	5.5	2.8	1.9	5
19 (through 0500)	340	2.3	—	1.0	2.2	3
20 (from 0100)	[3]	[3]	4.9	3.2	2.5	4
21 (through 0500)	304		—	2.8	2.0	2
10	336	2.3 3.0	5.2	2.9	3.9	3
11 (through 0400)	332	2.8	—	3.5	2.5	2
26 (from 2100)	—	—	—	1.8	2.4	1
27 (through 0800)	330	2.1	—	3.1	1.8	3
For comparison: average for undisturbed days	299	2.79	5.83	1.5	1.5	0

[1] Average for 12 hours of American magnetic K figure determined by seven observatories, on an arbitrary scale of 0 to 9, 9 representing the most severe disturbance.
[2] An estimate of the ionospheric storminess at Washington, on an arbitrary scale of 0 to 9, 9 representing the most severe disturbance.
[3] Recorder not in operation.

* Decimal classification: R113.61. Original manuscript received by the Institute, July 11, 1941. These reports have appeared monthly in the Proceedings starting in vol. 25, September, 1937. See also vol. 25, pp. 823–840: July, 1937. Report prepared by N. Smith, T. R. Gilliland, and C. O. Marsh.

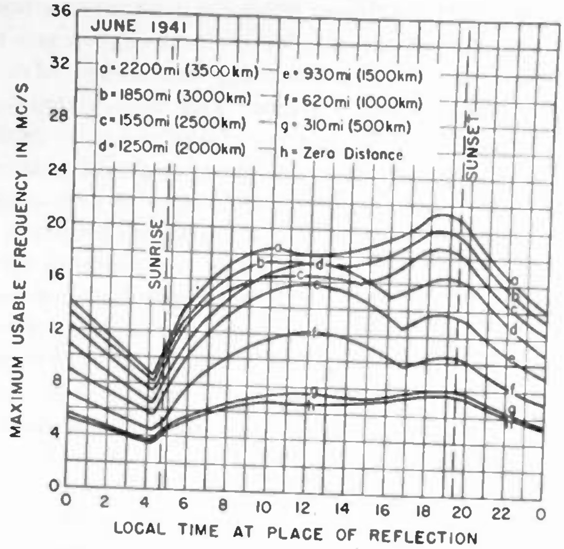


Fig. 1—Maximum usable frequencies for dependable radio transmission via the regular layers, average for undisturbed days, for June, 1941. The values shown were considerably exceeded during irregular periods by reflections from clouds of sporadic E layer (see Table III). These curves and those of Fig. 2 also give skip distances, since the maximum usable frequency for a given distance is the frequency for which that distance is the skip distance.

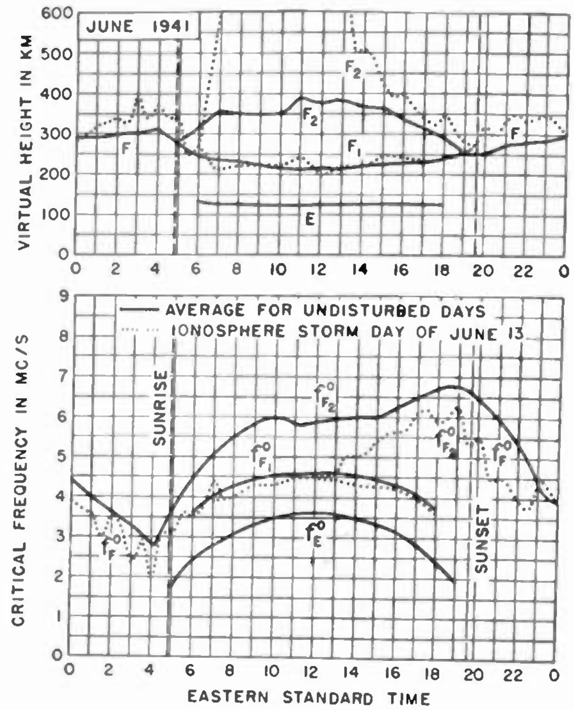


Fig. 3—Virtual heights and critical frequencies of the ionospheric layers, observed at Washington, D. C., June, 1941.

Critical frequencies for each day of the month are given in Fig. 4.

Ionospheric storms are listed in Table I. Beginning with last month's report, the scales used for the ionospheric and magnetic character figures run from 0 to 9 instead of 0 to 2 as previously. The magnetic-character figures given in Table I are averages, for each Greenwich half day, of the 3-hour magnetic K figures determined by the seven American-operated magnetic observatories.

The ionospheric storms during June were not as severe or frequent as in previous months. The details of the ionospheric storm day of June 13 are shown in Fig. 3. The open circles in Fig. 4 indicate the noon and midnight critical frequencies observed during the ionospheric storms listed in Table I. The sizes of the

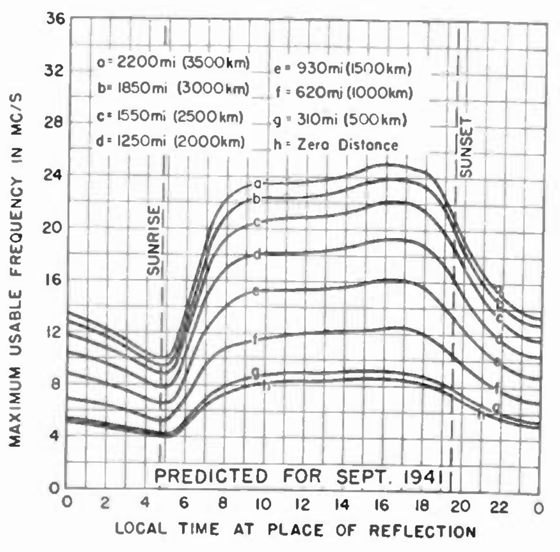


Fig. 2—Predicted maximum usable frequencies for dependable radio transmission via the regular layers, average for undisturbed days, for September, 1941. The values shown will be considerably exceeded during irregular periods by reflections from clouds of sporadic E layer. For information on use in practical radio-transmission problems, see pamphlets "Radio Transmission and the Ionosphere" and "Distance Ranges of Radio Waves" obtainable from the Natural Bureau of Standards, Washington, D. C., on request.

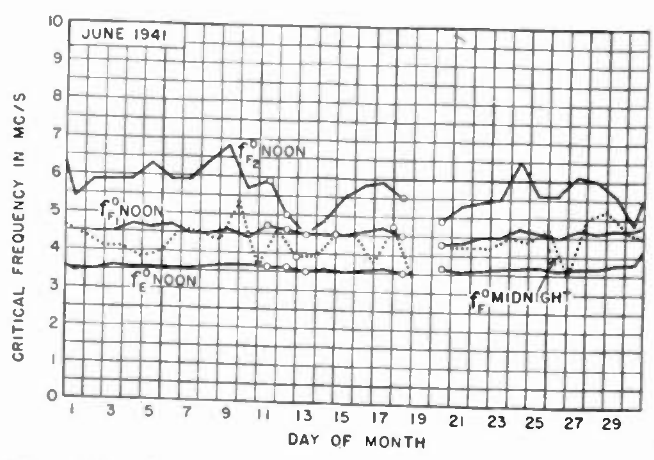


Fig. 4—Midnight $f_F{}^0$ and noon $f_E{}^0$, $f_{F_1}{}^0$, and $f_{F_2}{}^0$, for each day of June. Open circles indicate critical frequencies observed during ionospheric storms; sizes of circles represent approximate severity of storms.

circles roughly represent the severity of the storms.

Sudden ionospheric disturbances are listed in Table II. Table III gives the approximate maximum usable frequencies for good radio transmission via sporadic-E reflections.

TABLE II

SUDDEN IONOSPHERIC DISTURBANCES

Day	G.M.T. Beginning	G.M.T. End	Locations of transmitters[1]	Relative intensity at minimum[2]	Other phenomena
June 3	1138	1240	Ont.	0.0	Terr. Mag. Pulse[3] 1133–1150
4	1744	1840	Ont., D.C.	0.0	Terr. Mag. Pulse[3] 1743–1845
5	1431	1441	Ont., D.C.	0.01	Terr. Mag. Pulse[3] 1428–1435
6	1620	1730	Ont., D.C.	0.0	
9	1813	1840	Ont., D.C.	0.01	
14	1312	1350	Ohio, Ont., D.C.	0.0	
30	1355	1410	Ohio, Ont., D.C.	0.02	

[1] W8XAL, Mason, Ohio, was not recorded until June 11.

[2] Ratio of received field intensity during fade-out to average field intensity before and after, for station CFRX, 6070 kilocycles, 600 kilometers distant.

[3] As observed on Cheltenham magnetogram of the United States Coast and Geodetic Survey.

TABLE III

APPROXIMATE MAXIMUM USABLE FREQUENCIES IN MEGACYCLES, FOR RADIO TRANSMISSION VIA STRONG SPORADIC-E REFLECTIONS

Day	Hour, E.S.T. 00	01	02	03	04	05	06	07	08	09	10	11	12	13	14	15	16	17	18	19	20	21	22	23
June 1					19			19				23	34			23	27	19	16	16	15			
2								21	23	28	24	35	43											13
3	14	16		19	24	22	19												19	19	17			
4							24	23									25	28	25	23	19	19	20	17
5		17	16		16		21	24	23						22	27	27	24	19	22	39	34	19	
6		17	14	15	15	17	23	21										19	18	18	23	25	19	
7																				27				
8		15												34	36	28	22		24	18			15	19
9	17	19	19				18	23	25									25	35	32	41	45	33	50
10	50	34									25									23				
11	19	23	34	40	37	20	20	19	45	25										21	19			
12											20	22							18	23	22	21	20	
13		18												22										
14				19	27					45	46	45	50	50	34									
15							22	29	27	27									19	22	20	18		
16															27	22								22
17	19	14	24	19	15						28			36	23									
18							19	50	40	40	40				37	37		20	25	50	50	19		34
19	35	35	38	39	36	36	26																	
20											22													
21																			19	19	16			
22				15	15	17						25												
23																		27	42	34	24	19	14	
24	19	16	17					25	25	24										20	18			
25								44	27	24	50	37		28	34			22	45	22				
26	15	15	17	16				23	24	34	38	50									14	21	18	14
27	16																	23	18	19	39	50	50	23
28	15	18	19	21	40			23	23	34				39	34		35	24	42	50	45	37	28	24
29	17								23	45	35						34				18			
30	23	34	20	24	18	15	21		23															

Correspondence

Piezoelectric Crystals

I have read with interest the correspondence by Messrs. W. P. Mason and G. W. Willard in the September, 1940, issue of the PROCEEDINGS under the title "Piezoelectric Crystals."

This correspondence refers in some detail to a publication by the National Physical Laboratory entitled "Notation for Piezo-Electric Quartz." This pamphlet was drawn up by representatives of four British organizations interested in the subject and was also circulated, prior to publication, to some prominent workers in this field in France, Japan, and the United States of America. The notation suggested therein was considered to be helpful by all these workers in the field of piezoelectricity and it was hoped that the report would prove useful to other investigators.

The committee which was responsible for drawing up the report published by the Laboratory has been dissolved, and although Mr. Vigoureux is no longer a member of the staff of the Laboratory, I have had an opportunity of discussing with him, by correspondence, the proposals of Messrs. Mason and Willard. The following comments on their suggestions is the result of this correspondence.

Messrs. Mason and Willard write that the facet definition of "handedness" does not agree with the definition based on optical rotation as given in Vigoureux's books,[1,2] but nowhere in the books is it stated that the sense of the rotation applies to "an observer located at the light source." The usual convention is just the opposite, as Van Dyke points out,[3] and when it is adopted, the two definitions are in agreement.

The axial notation recommended in the National Physical Laboratory pamphlet is that adopted in Voigt's classical treatise, and followed by German writers and most English and American writers; some Japanese, and, as it now appears, some American writers advocate the use of right- and left-handed systems of axes for right- and left-handed quartz respectively. In practice there is little difference between the two procedures since in mathematical investigations right-handed axes presumably would always be employed. The results are applied to left-handed quartz by changing the system of axes in one case, in the other by reversing the signs of the piezoelectric constants and moduli.

It is immaterial whether the letters λ, μ, ν or l, m, n, be used for direction cosines, the former were recommended to avoid confusion with the m faces of the crystals, but no doubt this recommendation would be modified if there proved to be a strong current of opinion against it. Similar considerations would presumably apply to Π_{xx}, ξ_{xx}, etc. versus X_x, x_x, etc. for stress and strain.

The National Physical Laboratory recommendations state (last two paragraphs of page 8) that three angles, not two, are required for the specification of rectangular plates.

It is agreed that the many different and sometimes conflicting notations adopted by writers cause unnecessary difficulties to the reader of periodical literature on quartz, and tend to slow down progress in the subject, and in fact the National Physical Laboratory pamphlet was the outcome of a desire to bring the matters of standardization under discussion by an international body. The collaborators to the pamphlet did not regard their recommendations as final, but rather as forming the basis for a discussion of wider scope, in which their colleagues overseas would take part. The formation of the committee advocated by Mason and Willard is therefore welcomed, but it is felt that its findings should be given the form of proposals rather than decisions, for subsequent international discussion when circumstances permit.

R. L. SMITH-ROSE
Superintendent
Radio Department
National Physical Laboratory

[1] "Quartz Resonators and Oscillators," His Majesty's Stationery Office, 1931.

[2] "Quartz Oscillators," His Majesty's Stationery Office, 1939.

[3] Karl S. Van Dyke, "On the right- and left-handedness of quartz and its relation to elastic and other properties," PROC. I.R.E., vol. 28, pp. 399–406; September, 1940.

Institute News and Radio Notes

Address of President Terman at the Summer Convention Banquet Detroit, Michigan, June 24, 1941

Our Convention banquet tonight takes place at a crucial moment in our national history, in a city teeming with industrial activity directed towards national defense. I am sure that you have all shared with me the thrill of being immersed, even if only for a few days, in this "home town" of the internal-combustion engine, the many uses of which in these days are being made vivid as reports come in from abroad and as our own defense production accelerates. As I listened to the technical sessions at this Convention, I came to the feeling that the ultimate effect of radio techniques upon the national defense is fully as important as that of the internal-combustion engine.

Time was when radio was simply a form of telegraphy and telephony without wires. In the past few years, however, its by-product developments have mushroomed out to an extent that those pioneers who first worked in this field of electrical engineering can now hardly believe their recollections of the simplicity of its beginnings. It is one of the evidences of the rapid growth of radio that many of its pioneers are still among us. As engineering societies go, the Institute of Radio Engineers is among the youngest, the most virile, and the least fettered by tradition. Founded in 1912 to advance the art and science of radio communication through the presentation and publication of technical papers, your society's growth, membership, and influence have been contemporary with, and in considerable measure responsible for, the amazing expansion of the radio and electronic arts since the early twenties. There is no professional engineering society in the world which has the degree of authority and extent of coverage in the radio field that distinguishes the Institute of Radio Engineers. Among your past presidents and your medalists you cherish the greatest names associcated with the science and development of radio. You, of this Institute, have every reason for taking collective pride in the achievements of the individuals, past and present, who compose your illustrious roster.

In 1941 radio progress continues at accelerated pace. The method of frequency modulation of ultra-high frequencies is finding application in meeting many problems, and, in broadcasting, had its commercial debut at the beginning of the year. By midyear, television, which had met with some regulatory obstacles, will be off to a commercial start in cities scattered throughout the United States. Broadcasting, which heretofore has been denied sight, may be expected to respond to the stimulus of its new-found sense of vision. Either of these two developments by itself would, in normal times, represent a heavy load upon the technical forces of the industry. Coming in addition to the demands of national defense and faced with the consequent diversion of personnel, these new developments serve to emphasize what an extraordinary versatility of service to mankind the radio engineer is called upon to deliver at this time.

Today the armed services of the nation are drawing mightily upon the intellectual inheritance handed down by radio pioneers, and upon the current intellectual output of the membership of this Institute, a cross section of which forms this audience tonight. Radio and its allied arts will have much to do with the placing of force where force is needed in the event that this country goes to war. Some of these radio applications will be to the handling of communications, through whose agencies the armed forces of the nation will be supplied and moved. The swifter deployments of mechanized forces characteristic of war today make far greater demands upon telegraphic and telephonic communications than was the case in any preceding war. But in addition, electronic techniques are finding totally new employments—other than communication—in navigation and in searching out the enemy, whether he come by sea, land, or air. These applications are not simple forms of radio—they are among the most complex. Developments are being applied that were closed books to all but a dozen men a year ago. The most intricate military control equipment, much of it based upon radio devices, will be commonplace in our services when and if war comes to us.

And now we have an additional factor to complicate the problems of the radio engineer. Within recent weeks it has become evident that he faces a new challenge, and is called upon, like Israel of old, "to build bricks without straw." Confronted with shortages of certain strategic materials which up to now have been used freely in all radio equipment, we shall have to design many of our transmitters, receivers, and so on, with substitute materials and by new methods. Members of this Institute, leaders of their profession, will have to assume a *new* leadership in reducing the needs for standard materials, not only by substitution but

COMING MEETINGS

Pacific Coast Convention
Seattle, Washington
August 20, 21, and 22, 1941

Rochester Fall Meeting
November 10, 11, and 12, 1941

perhaps by reducing the number of different types of apparatus used; perhaps by standardizing, to an extent not heretofore known, on parts and types which are retained. Broadcast receivers, which we shall have to continue to manufacture to satisfy the military demand for preservation of civilian morale and safety, may have to be built with fewer tubes and with materials now strange to manufacturers. We are giving over a session at this Convention tomorrow to a special consideration of means available to us to meet this temporary difficulty.

With all the pressure that is being brought to bear upon imaginative radio thought for the prosecution of aggressively defensive measures, it is incumbent upon us, as an Institute representing the welfare of all its members, to take whatever action that may be open to us to insure that in this hour of our country's emergency there shall be no cessation of effort, no diminution in the flow of ideas, no radical changes in the organizations *through* which, and the tools *with* which, we work. There is a feeling of regret abroad throughout the radio engineering field that a squabble should have developed involving the government and industry in the midst of our country's defense effort. This creates confusion at the very moment when its morale and general effectiveness should be maintained at concert pitch.

We engineers know, and the communication units of our military services know, that the major forward steps which have so far been made in the radio field have originated in commercial and university laboratories, where scientists, engineers, and inventive technicians have been willing to devote their lives and efforts to the improvement of the radio art.

Within recent months, these laboratories have greatly expanded their facilities, and new ones are under way. The products of such laboratories will be of inestimable value to the United States for military and commercial purposes, both in this time of national emergency and in the critical times of peace which may be expected to follow. Nothing must be permitted to discourage the planning, construction, and operation of such laboratories and allied activities. The morale and output of the research and development engineers who work in them must not be destroyed by law, regulation, industrial practice, or the conflicts which deny stability to radio technical and commercial operations, which follow economically unsound and technically unjustifiable lines of control, or which withdraw incentive for the origination of new ideas and better methods by interfering with the natural way of gaining progress through the urge of the competitive spirit.

The history of the development of radio broadcasting is one of the brightest stories in American industry. Conceived in the daring and vision of its pioneers, born of the technical skill of our engineers, nourished and fostered by American principles of free competition, broadcasting in this country expanded at a rapid rate and successfully established an unparalleled service to the people. Under the former Federal Radio Commission, established some fourteen years ago, a benign form of government regulation began gradually to be applied, designed primarily to allot available frequencies among the applicants, in conformity with natural laws and within the limitations imposed by necessary international agreements. It became necessary, also, to begin policing the radio-frequency spectrum, but the policies of the Federal Radio Commission were broadly based upon sound engineering standards. No heavy regulating hand was laid upon radio technical progress. Thus an excellent service developed which received rapid public acceptance. The Commission made no attempt to usurp the engineering prerogatives of designers of broadcast transmitters and antenna structures.

The development of radio has been so rapid that the wisdom of regulatory legislation has always seemed to be inadequate to the demands. When the present Federal Communications Commission succeeded the former Radio Commission, it no doubt appeared to be wise to give it leeway in some respects and to limit its powers in others, as covered in the Communications Act of 1934. However, the results upon the industry have not been uniformly satisfactory to the Commission, to legislators, to the public, or to the radio industry.

One tendency manifested by regulation was to specify the internal aspects of station design and control, and the methods and equipment whereby the sought-for external results, in the electrical field, should be obtained. To engineers it is evident that regulation of station performance should be altogether restricted to the specification of external performance of a station and that in no instance should tubes, transmitting arrangements or circuits, station apparatus, measuring equipment, or the like be rigidly specified.

The Federal Communications Commission has a competent engineering division. Frequently we have wished that the Commission would heed the advice of its own engineers. We of the Institute of course recognize that the regulation of radio must be based in part on considerations other than technical. But these other considerations are not fundamental; they change from day to day and are subject to adjustment and treatment as occasion requires. On the other hand, technical requirements must always be met and never violated, or difficulty ensues—for technical requirements are based upon the laws of nature and cannot be disregarded with success. Consequently, we strongly feel that correct technical decisions are fundamental to any sound regulatory policy.

Nature determines how far radio waves travel and how strong they will be when they get there, and no amount of political gerrymandering will give good service to listeners if nature's laws are violated.

So much for our views of the situation. What should we do about it? Apparently the Institute will have to

do what everyone else is doing—go to Washington with its story. There we must make our collective voice heard in advocating the application of sound engineering principles, and, in the proper places and at appropriate times, urge legislation and regulatory policies consistent with such principles. I believe our position should be, first, that the interest, convenience, and necessity of the public obviously are best served by adopting technically correct and economically sound bases for regulation, rather than by major consideration of political situations, or of sectional and commercial rivalries; second, that the interests of the public and of radio engineers in the regulation of radio are identical, because of the fact that radio can continue to grow as a public service only so long as it serves the public well.

To give effect to these things which I have described tonight, the Board of Directors of the Institute has arranged an opportunity for you to vote tomorrow morning, by Yes or No ballot, on the following question:

> "Governmental public hearings affecting the radio industry are held from time to time. It has been several years since the Institute of Radio Engineers has actively participated in such hearings. With respect to future hearings, do you favor the Institute's appearing for the purpose of presenting the engineering view on the subjects under consideration?"

I hope that there may be free discussion of this matter and a decisive vote.

Hearings are now being held before the Senate Interstate Commerce Committee which indicate that the Commission, Congress, and the industry are unhappy about something. Perhaps this is an indication that the Communications Act of 1934 is out of date and does not now meet the requirements of a greatly advanced radio art. As we all know, much progress has been made in all phases of radio in these past seven years and it is too frequently the case that legislation falls behind the rapid progress made in an industry which it was designed to regulate. For example, the Radio Act of 1927 became outmoded after seven years of experience, which condition was recognized by Congress when it passed the Communications Act of 1934. It is thought that herein lies the seat of the troubles that now beset the radio industry. The Institute believes that these difficulties could be eliminated to a large degree if Congress would recognize this fact and would consider the drafting of a new act. Perhaps this would lead to other hearings designed to obtain the views of all interested parties. At that time there should be present an opportunity for the Institute to assist in formulating such legislation as will encourage progress and assure improvement in the radio services of our Country.

These matters which I have discussed represent the outstanding problems confronting radio engineers today, and, the opportunities for service to their profession and to their country. We can take heart in facing the problems from the fact that the opportunities for service are so great and that the importance of our work is now becoming so widely recognized. I look forward to a future for our profession even brighter than its past.

Regional Directors Proposed in New Constitutional Amendment

Plan

Members of the Institute will soon be asked to vote upon a series of amendments to the Institute Constitution. These create the new office of regional Director, and in addition include minor modifications that correct small imperfections in the present Constitution and make the new Executive Committee setup fully effective.

The amendments relating to the regional Directors were approved by the Board of Directors at their April meeting, and are given in full in the report of that meeting appearing on page 138 of the March issue of the Proceedings. These provide for the establishment of geographical areas, not to exceed eight in number, each of which shall have a regional Director nominated by petition from the region, elected by the voting members of the region, and holding office for two years.

If the proposal for regional Directors is approved, it is anticipated, on the basis of discussion in the Board, that each regional Director would be charged with the responsibility of looking after the welfare of the Institute in his area. This would include: (1) representing his own constituency on the Board of Directors so that its views, suggestions, and criticisms could be effectively presented to the governing body and to the officials of the Institute; (2) assisting and co-ordinating the activity of Sections and other Institute groups in the area for the purpose of strengthening the Institute; (3) keeping his own constituency properly informed about Institute policies; and (4) settling differences that arise between various groups of members in the area, or between members in the area and Institute headquarters. Each regional Director would be expected to submit a written report at the end of every year covering conditions in his area, changes that had taken place during the year, and suggestions for the future.

The plan for regional Directors as submitted to the Board for consideration assumed that bylaws would be passed providing for two specially designated Board of Directors' meetings each year, preferably associated with Institute conventions, which the regional Directors would be expected to attend, and for which they would receive a traveling allowance approximating railroad fare plus Pullman. These meetings would have

an especially prepared agenda providing for broad discussion of Institute policies and problems. Regional Directors, as *ex officio* members of the Board, would in addition be entitled to attend other Board meetings (but without traveling allowance), and it is expected that a majority would be able to attend more than two Board meetings each year.

The cost of bringing the regional Directors to New York twice a year would approximate $800 to $900 on the basis of the *actual cost* of first-class railroad fare plus Pullman, assuming eight directors distributed in a reasonable manner over the United States and Canada. This is approximately 1.25 per cent of the total income of the Institute for 1941.

The constitutional amendments providing for regional Directors were discussed at considerable length by the Board, and while they have the approval of a large majority of that body, there were several who doubted the wisdom of making such an innovation. Accordingly, the Board has arranged for the preparation of arguments pro and con, which are given below. Every voting member of the Institute should study these arguments carefully, decide upon what appears to be the proper policy for the Institute to follow, and then return his marked ballot promptly after it arrives.

Frederick Emmons Terman, *President*

Discussion

The regional Director plan is the first concrete proposal which gives promise of solving the difficulty which has heretofore beset the Board of Directors, in securing adequate representation of the Institute's widespread membership. With the advance of time, it has become increasingly difficult to secure exact and prompt information on the views of the more distant parts of the membership, as well as to acquaint the membership with the problems and activities of the Board. The regional Director plan provides these needed contacts. Discussion of methods for securing regional representation has occurred in the Board from time to time but none of the suggestions made has resulted in any method as effective as this plan, particularly from the standpoint of securing a widespread geographical representation.

Adequate representation of any certain section of the country requires very specific allocations of duties and responsibilities to that end. It is obvious that a Director who resides in Massachusetts, for example, cannot be familiar with the views and problems in Illinois; therefore, Directors at large do not provide the needed intimate two-way communication channel. The regional Director plan meets this need by insuring wider participation in the management of the Institute and better contacts between its members and the governing body. In the last fifteen years there has been only one Director (who served for one year) who lived farther than 300 miles airline from New York City at the time of election. The plan for regional Directors will do much to remedy this unhealthy situation.

The plan for regional Directors provides against the natural condition that useful work does not get accomplished unless somebody has definite responsibility therefor. As the Institute is now organized, there is no one on the governing body who has a definite responsibility for seeing that the Institute functions properly in any particular area. The Board of Directors needs a specialist from each area of the Institute, who will be responsible for the welfare of the Institute in that area, who will keep the Board informed as to the problems of the area, and who will serve as a combined good-will ambassador, trouble-shooter, co-ordinator of regional activities, and promoter for the Institute.

Regional participation in the management of a technical society is neither new nor radical, and is used successfully by other engineering and scientific societies, both large and small. Some of these organizations consider such representation sufficiently important to expend large sums of money for this purpose.

The cost of the plan as proposed is small, about $100 per region per year which would cover first-class railroad tickets plus Pullman accommodations for two meetings a year. Inasmuch as the cost will be limited to the figure established by the bylaws which will implement the plan as it is set up, expenditures for this purpose can never be increased except by changes in the bylaws which require a two thirds vote of the Board.

The plan requires regional Directors to attend at least two meetings per year. Therefore, the objects of the plan are certain to be accomplished. It is true that other Directors need not conform to such a requirement, but experience has shown that Directors living near New York City do not require constitutional compulsion or financial assistance to assure frequent attendance at meetings. The provision of these two specified meetings per year permits preparation in advance for presentation and discussion of matters of wide and basic interest or which otherwise might not come to the attention of the Board in the absence of regional representation.

The regional Director plan will remove the lack of close contact with headquarters, which lack was disclosed by the reports of two of our Presidents who recently made national tours among our various Sections, and will ensure continuous and efficient contact.

The regional Director plan is workable, practical, and effective. It provides contact between individual members and the management of the Institute through local representatives of their own choosing, and it brings to the management specialists familiar with the individual regions. It places in a definite and workable way responsibility for the proper functioning of the Institute in each individual area. These things are all desirable, and are now lacking.

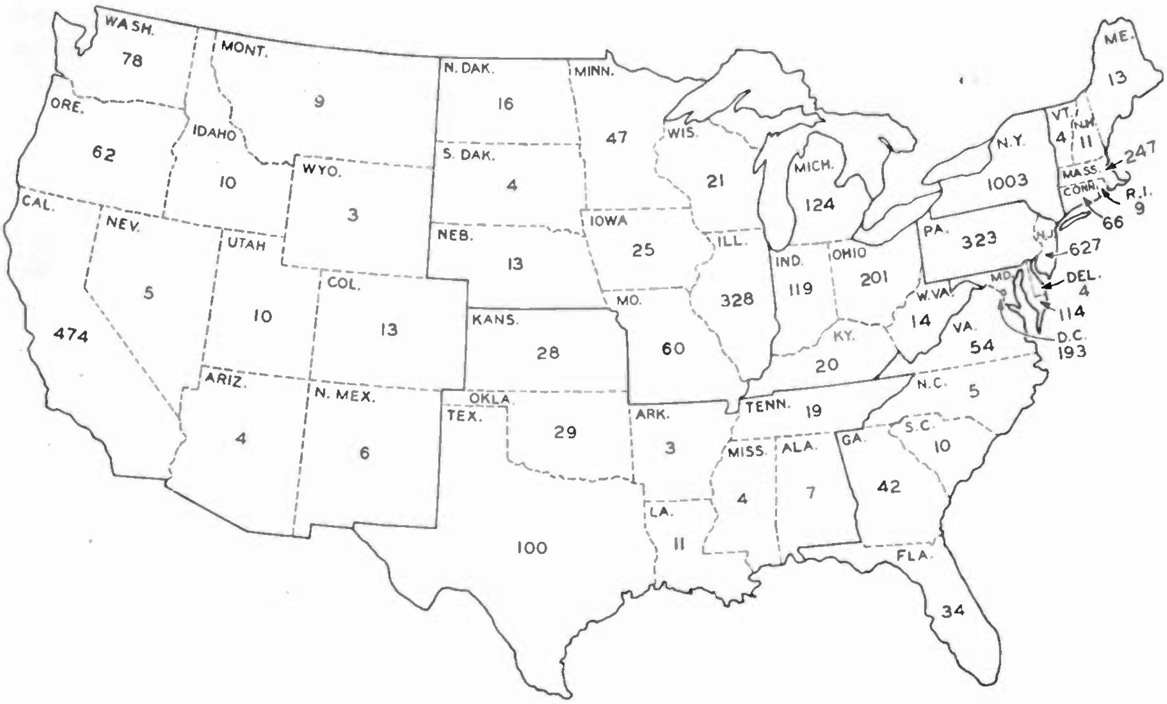
C. M. Jansky, Jr.
Haraden Pratt
A. F. Van Dyck

Regional Directors are proposed as a method of obtaining closer relations between the membership and the Board of Directors, the prime requirement being a better interchange of knowledge. If there is a need for this, the effectiveness and cost of any plan should be examined and compared with other ways of doing the job. We believe that regional Directors will provide neither the most effective answer nor even the best answer in relation to its cost and submit the following information and suggestions.

Management of the Institute. Since the founding of the Institute, the Board of Directors has grown from nine to twenty-one members. Regional Directors would raise this number to a possible twenty-nine. It has already been recognized that twenty-one members are too many and, recently, the detailed management of the Institute has been delegated to an Executive Committee of seven Board members. Even with the Executive Committee, the Board will continue to meet ten or more times yearly to direct the business of the Institute.

However, with a large Board, and one having a quarter of its members able to attend only a couple of meetings each year, it will be a logical step to put all Directors on the same basis by holding only a few meetings to each of which the expenses of all Board members will be paid. This offers the additional incentive of a still broader geographical distribution of the Board membership. While such a system has an attractive appearance, it would permit a majority of a small Executive Committee to run the organization with impunity as all Board members will be responsible for their acts. The membership cannot protect itself against such a system even with the power of the ballot.

Cost of Regional Directors. The accompanying map shows the number of members in each state at the end of 1940. Eight regions have been outlined to represent one possible arrangement, no official action in the matter having been taken. It may be assumed, for purposes of computing costs, that the regional Director will come from the largest section in each region. The following table gives data on the number of members in each region, the city from which the Director will travel, its distance from New York City, and the cost at $8\frac{1}{2}$ cents per mile, one way, which figure has been used in Board discussions as representative of what a similar organization allows.



Region	*Members*	*Director's City*	*Miles from New York City*	*8.5¢/mile*
Western	674	Los Angeles	3111	$264.44
North Central	514	Chicago	909	77.27
South Central	201	Dallas	1739	147.82
Great Lakes	478	Detroit	648	55.08
Pa.-N. J.	950	Philadelphia	92	7.82
New York	1003	New York City	—	—
New England	350	Boston	232	19.72
South Atlantic	456	Washington, D. C.	227	19.30
	4626		6958	$591.45

The proposal assumes attendance at two meetings each year so a total expenditure of about $1200 will be required.

What $1200 Buys. For $1200 per year, a number of things might be purchased, or, conversely, might have to be given up if additional funds are not available. Some of these things are:

a) About 60 Proceedings pages, equivalent roughly to one issue, or

b) Two major standards reports, or
c) Thirty per cent of the rebates paid to sections for their operation, or
d) A visit by the President and one other officer each year to sections in the United States and Canada, or
e) One clerical worker. The Secretary's staff has been undermanned for several years and still lacks at least two executive and two clerical workers.

ALTERNATE PROPOSAL. At the end of 1940 there were 20 Sections in the United States and 2435 of our 4626 domestic members (53 per cent) were located in them. In New York City and its immediate environs there were about 1275 members (27 per cent) who presumably are adequately represented at the present time. These 20 Sections and New York account for 80 per cent of the United States membership. At the end of 1940, 46 per cent of the domestic membership was located in New York or in Sections in which members of the Board of Directors resided.

A mechanism designed to utilize the existing Section structure could readily be established at practically no cost. The Executive Committee of a Section undoubtedly knows more about the problems of the members of a Section and can be more effective in distributing information to them than any single Director charged with caring for a territory comprising thousands of square miles and which includes from two to four Sections.

For many years, the Sections Committee has met annually at conventions to discuss purely Section problems. About half of the Sections are represented at these meetings. Such meetings, attended also by the Board of Directors, could be held at each of our two yearly conventions to discuss all Institute matters and thus provide effective participation by the Sections in the management of the Institute.

YOUR INSTITUTE AND YOUR MONEY. This is your Institute and your money. If you feel that you need this proposed additional representation and that it will improve the management of the Institute, it is your duty to vote for it. If, on the other hand, you think the existing form of management has provided reasonable and adequate government, or that the alternate Section method of broadening the views of the Board would be successful, perhaps you would prefer some of the other things that your money will buy.

AUSTIN BAILEY
H. M. TURNER
H. P. WESTMAN

Contributors

Murray G. Crosby (A'25–M'38) was born at Elroy, Wisconsin, on September 17, 1903. He attended the University of Wisconsin from 1921 to 1925, and received the B.S. degree in electrical engineering in 1927. From 1925 to 1927 he was with the Radio Corporation of America, and from

M. G. CROSBY

A. W. MELLOH

1927 to date he has been with R.C.A. Communications, Inc.

❖

Arthur W. Melloh (A'33) was born at Wrenshall, Minnesota, on December 8, 1907. In 1932 he received the B.E.E. degree from the University of Minnesota and from 1932 to 1933 he was engineer at XENT, Nuevo Laredo, Mexico. From 1934 to 1935 he was an instructor at the Dodge Radio Institute and in 1936 Mr. Melloh returned to the University of Minnesota from which he received the M.S. degree in electrical engineering in

M. E. STRIEBY

1937 and the Ph.D. degree in 1940. From 1937 to 1940 he was an instructor in the department of electrical engineering at the University of Minnesota and since 1940 has been with the Associated Electric Laboratories, Chicago, Illinois. He is a member of Sigma Xi and an associate member of the American Institute of Electrical Engineers.

M. E. Strieby (M'31) was born in 1893 in Colorado Springs, Colorado. He received the degrees of bachelor of arts in 1914 from Colorado College, bachelor of science in 1916 from Harvard University, and bachelor of science in electrical engineering in 1916 from the Massachusetts Institute of Technology. During the year following his graduation from the Massachusetts Institute of Technology he was employed as an engineer for the New York Telephone Company, and during 1917–1919 he served as captain in the Signal

W. O. Swinyard

Corps of the United States Army, American Expeditionary Forces. Mr. Strieby was employed by the American Telephone and Telegraph Company, in 1919, and became engaged in research on the transmission features of telephone repeaters and associated equipment. In 1929 he was transferred to the technical staff of the Bell Telephone Laboratories where he was occupied with studies of new high-frequency carrier apparatus and technique. In particular he was responsible, as carrier transmission research engineer, for the development of coaxial cable systems. Studies of means for the transmission of television over such systems and over shorter wire lines in cities were also under his supervision. In December, 1940, Mr. Strieby was transferred to the American Telephone and Telegraph Company to become engineer of transmission for the Long Lines department. He is a Fellow of the American Institute of Electrical Engineers.

William O. Swinyard (A'37–M'39) was born July 17, 1904, at Logan, Utah. He received the B.S. degree in physics from the Utah State Agricultural College in 1927 and was an instructor in mathematics there in the summer of 1927. From 1927 to 1930 he was an instructor in physics,

C. L. Weis

mathematics, and music in secondary schools in southern Idaho. After graduation from the RCA Institutes in 1930, he was employed in the Bayside laboratory of the Hazeltine Service Corporation. In 1934 he was transferred to the New York laboratory. He enrolled in the graduate school of Columbia University in 1934 and during the following two years pursued evening courses in vacuum tubes and communications networks. Since 1937 he has been an engineer in the Chicago laboratory of the Hazeltine Service Corporation. He is a member of the Radio Club of America, the Acoustical Society of America, and the Illinois Professional Communications Engineers Association.

C. L. Weis (A'39) was born January 17, 1901 at La Crosse, Wisconsin. He received the B.S. degree in electrical engineering from the Massachusetts Institute of Technology in 1922. He joined the engineering department of the Western Electric Company upon graduation. Mr. Weis is now with the Bell Telephone Laboratories engaged in development work on television apparatus and systems for use over wire lines of the American Telephone and Telegraph Company. During the past few years his work has been specifically the application of such circuits to broad-band cables of the coaxial type. He has supervised all the long-distance television transmissions

E. S. Winlund

over coaxial cables which have been made in this country.

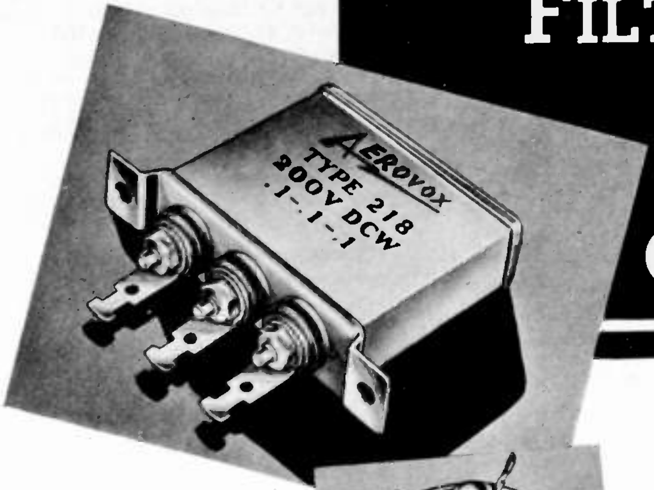
E. S. Winlund was born in Oakland, California, in November, 1912. He was graduated from the University of California with the B.S. degree in electrical engineering communications in 1936, and then entered Massachusetts Institute of Technology as a graduate student cooperative with the General Electric Company. In January, 1937, he entered the regular "Test" course in Schenectady where his assignments included government radio transmitters, thyratrons, motors and generators, industrial control, publicity department, and television tube development. In January, 1939, Mr. Winlund joined the broadcast transmitter engineering staff of the RCA Manufacturing Company at Camden, specializing in frequency-modulation development and design. In December, 1940, he transferred

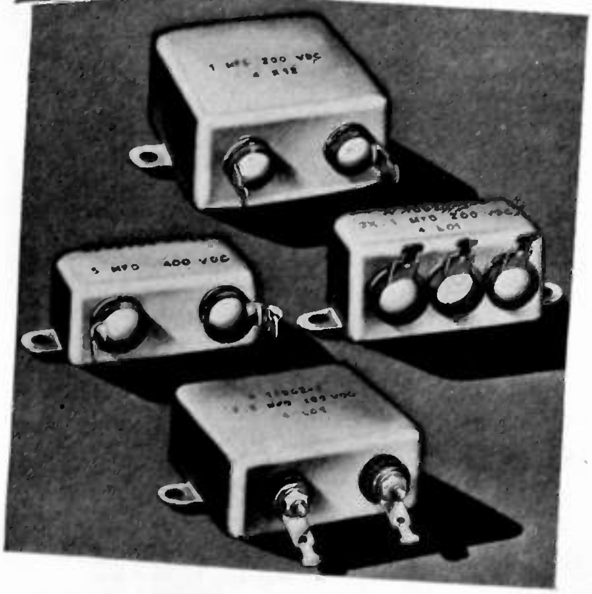
S. K. Wolf

to the broadcast transmitter sales division in the Camden office.

Sidney Kellum Wolf (A'40), was born at Baton Rouge, Louisiana, on August 29, 1901. He received the B.A. degree from Saint Stanislaus College; the B.S. degree from Louisiana State University in 1923; and the M.S. degree from the Sheffield Scientific School, Yale University, in 1926. From 1922 to 1923 Mr. Wolf was an engineer with the Westinghouse Electric and Manufacturing Company in Pittsburgh; 1923 to 1928 an instructor at Sheffield Scientific School, Yale University; 1928 to 1938, director of acoustic engineering at Electrical Research Products, Inc., New York City; 1937 to 1938, district manager of Erpi Picture Consultants, Inc. Since 1938, Mr. Wolf has been president of Acoustic Consultants, Inc., and New York representative of The Brush Development Company since 1939. He was an American representative at the International Electrical Congress, Paris, 1932, and representative of the American Standards Association at the International Standards Association Congress, Budapest, 1937. He is a director of Wolf's Bakery, Inc., and a director of the Edison Foundation. Mr. Wolf is a member of Tau Beta Pi, and Theta Xi.

FILTER and BY-PASS *Oil-filled* CAPACITORS





● For filtering and by-passing functions, AEROVOX offers both Hyvol and mineral oil capacitors, as well as wax-impregnated wax-filled capacitors, in a wide choice of casings, mountings, terminals. And AEROVOX stresses its Hyvol capacitors particularly for use in equipment subject to sub-zero operating conditions. Such capacitors are considerably more stable with temperature variations than are those using synthetic oils. Hyvol capacitors show no appreciable drop in capacity until temperatures of −20° F. (−29° C.) are reached. At −40° F. (−40° C.) the maximum capacity drop that can be expected is of the order of 5 to 10%. Specify Hyvol, therefore, when dealing with a wide range of temperatures.

RECTANGULAR-CAN SERIES 16 & 18

Designed for aircraft and marine radio assemblies where most efficient use of space and minimum weight are essential. Oil-impregnated sections in rectangular corrosion-proof metal containers filled with oil and hermetically sealed by soldering. Special immersion-proof terminals. With top terminals (Series —16) and bottom terminals (Series —18). Standard ratings: 200, 400, 600 and 1000 v. D.C.W. .01 to 1.0 mfd. Also double and triple sections.

DRAWN-CASE SERIES 30

Non-inductive oil-impregnated paper section encased in one-piece drawn metal case with soldered bottom plate for hermetic sealing. Terminals of double rubber bakelite for absolutely immersion-proof terminal assembly. Terminals normally on side of case, but can be placed on both sides, on top or on bottom. Standard ratings: 200, 400 and 600 v. D.C.W. .05 to 2.0 mfd. Also double and triple sections.

ROUND-CASE SERIES 40

Aluminum-can unit. Molded bakelite top. Soldering lug terminals riveted to studs molded in cap. Immersion proof. Mounting ring furnished. Standard ratings: 200, 400 and 600 v. D.C.W. Capacities of .05 to 2.0 mfd. Also double and triple sections.

Ask for Copy . . . *Send Your Problem . . .*

● If you are engaged in designing, building or maintaining commercial radio or electronic equipment, write on business letterhead for copy of Transmitting Capacitor Catalog. Otherwise ask our local jobber for latest radio catalog. Send your capacitance problem for our engineering aid.

AEROVOX CORPORATION
New Bedford, Mass.
Sales Offices in All Principal Cities

Booklets, Catalogs and Pamphlets

The following commercial literature has been received by the Institute.

CERAMICS • • • *American Lava Corporation, Testing Laboratory, Chattanooga, Tennessee. Engineering Data and Property Chart No. 416, 2 pages, $11\frac{3}{4} \times 21$ inches.* A tabular and graphical summary of the electrical and mechanical properties of the principal bodies manufactured by this company.

POWER SUPPLIES • • • *Standard Transformer Corporation, Chicago, Illinois. Catalog 109-D, 12 pages, $8\frac{1}{2} \times 11$ inches.* Specifications on power packs "for all power change purposes."

RESISTORS • • *Ohmite Manufacturing Company, 4835 West Flournoy Street, Chicago, Illinois. Catalog Number 17, 12 pages, $8\frac{1}{2} \times 11$ inches.* Fixed and adjustable resistors, principally of the vitreous-enamel type.

MICROPHONES • • • *American Microphone Company, Ltd., 1915 South Western Avenue, Los Angeles, California. Catalog No. 37, 8 pages, $8\frac{1}{2} \times 11$ inches.* Crystal and dynamic microphones.

LOUD SPEAKERS • • • *Jensen Radio Manufacturing Company, 8601 South Laramie Avenue, Chicago, Illinois. Condensed Catalog No. 125, 8 pages, $8\frac{1}{2} \times 11$ inches.* Speakers, reproducers, and projectors.

SHIELDED CONDUCTORS • • • *American Steel and Wire Company, Cleveland, Ohio. Bulletin, 12 pages+cover, $8\frac{1}{2} \times 11$ inches.* Describes power cables "insulated" with a layer of semi-conducting rubber for shielding purposes.

RELAYS • • • *Standard Electrical Products Company, 417 First Avenue, Minneapolis, Minnesota. Catalog 641, 4 pages, $8\frac{1}{2} \times 11$ inches.* Relays of several types for application in communications equipment.

INSTRUMENTS • • • *Triplett Electrical Instrument Company, Bluffton, Ohio. Catalog A, 12 pages, $8\frac{1}{2} \times 11$ inches.* Electrical and mechanical specifications on panel-mounting electrical indicating instruments.

COMPONENTS • • • *American Phenolic Corporation, Chicago (Cicero Post Office), Illinois. Catalog No. 65, 42 pages, $8\frac{1}{2} \times 11$ inches.* A listing of cable connectors, sockets, coaxial cables, and polystyrene insulating material.

BRIDGES • • • *General Radio Company, 30 State Street, Cambridge, Massachusetts.* "The General Radio Experimenter," *July, 1941, 8 pages, 6×9 inches.* Gives instructions, with a discussion of the accuracies to be expected, for bridges made up of individual laboratory-type components.

AIRCRAFT DYNAMOTORS • • • *Pioneer Gen-E-Motor, Chicago, Illinois. Catalog 16 pages +cover, $8\frac{1}{2} \times 11$ inches.* Description and operating data on dynamotor-type power supplies for communication installation on aircraft.

CAPACITORS • • • *Aerovox Corporation, New Bedford, Mass.* The Aerovox research worker, *March 1941 and April 1941 4 pages $8\frac{1}{2} \times 11$ inches.* These two issues contain, respectively, articles on "Fixed Condensers in Radio Transmitters" and "R. F. Power Amplifier Operation."

(*Continued on page iv*)

(*Continued from page ii*)

BOOKS • • • *John Wiley and Sons, Inc., 440 Fourth Avenue, New York, New York. Catalog, 32 pages plus cover, 6¼×9¼ inches.* A list of this publisher's technical books of interest to personnel engaged on the national defense program. Includes several of interest to engineers.

FUSES • • • *Littelfuse Incorporated, 4757 Ravenswood Ave., Chicago, Illinois. Bulletin, 4 pages, 8½×11 inches.* Specifications on small fuses, fuse panels, and mercury switches. Particular emphasis on aircraft applications.

HOME-STUDY COURSE • • • *Smith Practical Radio Institute, 1311 Terminal Tower, Cleveland, Ohio. Information booklet, 32 pages+cover 3½×6 inches.* A description of a home-study course in radio and communications engineering.

TUBE DATA, (RCA) • • • *RCA Manufacturing Company, Inc., Harrison, N. J. Application Note No. 114, 2 pages, 8½×11 inches.* "Use of Cushioned Sockets in Small Receivers." *Application Note No. 115, 7 pages, 8½×11 inches.* "A Discussion of Noise in Portable Receivers." *Application Note No. 116, 6 pages, 8½×11 inches.* "Properties of Untuned R-F Amplifier Stages."

TUBE DATA (Ken-Rad) • • • *Ken-Rad Tube & Lamp Corporation, Owensboro, Ky. Data bulletin, 16 pages+cover, 8½×11 inches.* A bulletin giving "Essential Characteristics on Metal and Glass Radio Tubes."

VIBRATION ISOLATION • • • *Lord Manufacfacturing Company, Erie, Pennsylvania. Chart, 4 pages, 8½×11 inches.* This chart shows the percentage of vibration isolation it is possible to obtain with the bonded rubber mountings manufactured by this company.

Current Literature

New books of interest to engineers in radio and allied fields—from the publishers' announcements.

A copy of each book marked with an asterisk (*) has been submitted to the Editors for possible review in a future issue of the Proceedings of the I.R.E.

* ALIGNING PHILCO RECEIVERS (Volume II). BY JOHN F. RIDER. New York: John F. Rider Publisher, Inc., June, 1941. xv+192 pages, illustrated, 5×7¼ inches, cloth. $1.60.

* PHYSIK UND TECHNIK DES TONFILMS (Physics and Engineering of Sound Motion Pictures). BY HUGO LICHTE and ALBERT NARATH, Lecturer at the Technische Hochschule, Berlin. Leipzig: Verlag Von S. Hirzel, 1941. viii+371+11 index pages, illustrated, 7×10 inches, paper. rm. 26.

POSITIONS OPEN

The following positions of interest to I.R.E. members have been reported as open on August 8. Make your application in writing and address to the company mentioned or to

Box No. . . . . . .

PROCEEDINGS of the I.R.E.

330 West 42nd Street, New York, N.Y.

Please be sure that the envelope carries your name and address

AIR CORPS INSTRUCTORS

Competitive examinations are being conducted by the U. S. Civil Service Commission for Assistant Instructors, Junior Instructors, and Student Instructors in the Air Corps Technical Schools. Radio engineering and radio operating are among the optional branches of specialization.

Salaries offered are $2600, $2000, and $1620 per year, respectively. Applicants must have completed a high-school course or the equivalent and must have had certain specified amounts of experience against which specialized college training may be offered.

For full details consult the Commission's announcement 7-125 and a recently issued amendment. These can be obtained from the Secretary, Board of U. S. Civil Service Examiners, Chanute Field, Rantoul, Illinois; from the Commission in Washington, D.C.; or at any first- or second-class post office.

DEVELOPMENT ENGINEERS

The RCA License Laboratory, New York, has positions open for senior radio engineers having qualifications for development and consultation work. Good technical education, experience of 5 to 10 years, and U. S. citizenship required. Interesting work, broadening experience, and wide contacts. Write fully to Manager, RCA Laboratory, 711 Fifth Avenue, New York, N.Y.

PROCUREMENT INSPECTOR

The United States Civil Service Commission is announcing examinations for several grades of procurement inspector for the material division of the Air Corps, of the War Department. Salaries range from $1,620 a year for junior procurement inspector to $2,600 a year for senior procurement inspector.

Certain education and experience requirements must be met. In the radio branch, one of the eleven optional branches of specialization, experience must have been acquired on commercial radio-communications equipment.

For additional information and application forms write to the Secretary, Board of United States Civil Service Examiners, Wright Field—Fairfield Air Depot, Wright Field, Dayton, Ohio, or the Secretary, Board of United States Civil Service Examiners at any first- or second-class post office.

Attention Employers . . .

Announcements for "Positions Open" are accepted without charge from employers offering salaried employment of engineering grade to I.R.E. members. Please supply complete information and indicate which details should be treated as confidential. Address: "POSITIONS OPEN," Institute of Radio Engineers, 330 West 42nd Street, New York, N.Y.

The Institute reserves the right to refuse any announcement without giving a reason for the refusal

INDEX

DISPLAY ADVERTISERS

This Valuable

ENGINEERING DATA AND PROPERTY CHART

Is Yours on Request!

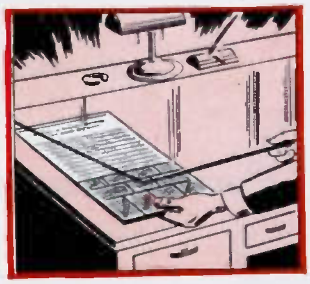
EVERY man who uses electrical insulating materials, chemical ceramics—in fact any ceramic parts for technical applications, and every research laboratory will find this Property Chart of value. It will be sent free on request.

The chart contains exact data on the physical properties of the most frequently used ceramic compositions of the American Lava Corporation. It gives the predominant physical characteristics such as: specific gravity, density, volume, water absorption, chemical resistance, color, softening temperature, resistance to heat (safe limit for constant temperature), hardness, linear co-efficient of thermal expansion, tensile strength, compressive strength, modulus of rupture, resistance to impact, thermal conductivity, dielectric strength, volume resistivity at various temperatures, Te values, dielectric constant, power factor, loss factor and capacity change per °C between 20-80° C. The chart also contains graphs which give valuable information on physical characteristics which are difficult to reduce to figures, such as linear thermal expansion, dielectric strength and volume resistivity at varying temperature.

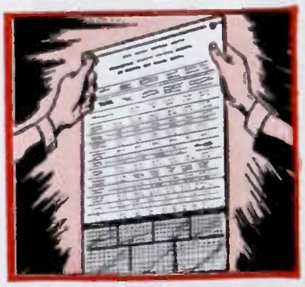
This chart is strictly an engineering tool. All data have been carefully compiled and checked over a long period of time.

The chart portrays most forcibly an outstanding advantage of ALSIMAG. In using ALSIMAG the engineer is not limited to one or two ceramic materials. ALSIMAG offers a wide variety of ceramic compositions of controlled and definitely determined physical characteristics for your application. ALSIMAG is economically custom made into size, shape and form indicated by your requirements.

Exactly file size, to fit in cabinet or desk file.



All engineering data can be seen thru desk glass.



Hangs on the wall of engineering department.

Please address requests for Property Chart to:

TESTING LABORATORY, AMERICAN LAVA CORPORATION, CHATTANOOGA, TENNESSEE

ALSIMAG FROM CERAMIC HEADQUARTERS

Trade Mark Reg. U.S. Pat. Off.

AMERICAN LAVA CORPORATION • CHATTANOOGA • TENNESSEE

CHICAGO • CLEVELAND • NEW YORK • ST LOUIS • LOS ANGELES • SAN FRANCISCO • BOSTON • PHILADELPHIA • WASHINGTON, D C

"Consider a Capacitor's Dependability"

THERE'S EXTRA DEPENDABILITY IN LONGER LASTING C-D CAPACITORS

Sound advice to engineers and purchasing agents is the capacitor recommendation of Canadian Colonial Airways. Consider dependability. You get *extra* dependability when you specify Cornell-Dubilier capacitors, because the broadest experience in the industry is behind them — more than 31 years of specialization in capacitor manufacturing.

Cornell-Dubilier superiority is hidden to the eye. But it's there in the capacitor ingredients . . . in the quality of parts, in the skillfulness of fabrication — and it shows in service. So much so that Cornell-Dubilier capacitors are the unanimous choice today of America's leading airlines for all aircraft and ground transmitting and receiving communications equipment. Whatever your capacitor requirements, it pays to specify C-Ds.

MICA · DYKANOL · PAPER · WET & DRY ELECTROLYTIC CAPACITORS

CAPACITORS MAY LOOK ALIKE BUT . . .

There is extra long life, extra uniformity and dependability built into C-Ds. Next time you specify capacitors look for the Cornell-Dubilier seal of experienced engineering. And get the hidden extras at no extra cost. Send for Catalog. Cornell-Dubilier Electric Corporation, 1012 Hamilton Blvd., South Plainfield, New Jersey.

Cornell Dubilier

SOUTH PLAINFIELD, N. J. NEW BEDFORD, MASS.

. . . MORE IN USE TODAY THAN ANY OTHER MAKE . . .

FOR ULTRA-HIGH FREQUENCIES . . .
A NEW OSCILLATOR with Good OUTPUT

A—*Counter and dial reading direct in approximate wavelength; used with conversion chart to read frequency*
B—*Approximate wavelength readings in centimeters*
C—*Conversion scale by means of which oscillator frequency can be determined to approximately ±2½ per cent*
D—*Short-circuiting piston which varies length of coaxial line and determines frequency*
E—*Lead screw which drives plunger "D"*
F—*W.E. Type 316-A Oscillator tube*

TYPE 757-P1 Power Supply

TO MEET the demand for a laboratory oscillator of good power output, with reasonably accurate frequency calibration, General Radio has developed and now announces the Type 757-A U-H-F Oscillator. This oscillator has adequate power output for laboratory and field-strength measurements at frequencies up to 600 megacycles. It uses a single tube which energizes a resonant coaxial line, the length of which determines the oscillator frequency. The effective length of the line is varied by means of a short-circuiting plunger which is driven by means of a lead screw. The output circuit is coupled to the main axis of the oscillator through a rod. This rod is so placed that the output impedance is approximately 75 ohms.

The oscillator is direct reading in approximate wavelength in centimeters. By means of a conversion-table scale permanently mounted on the oscillator housing, the resulting frequency in megacycles can be determined with an accuracy of approximately ±2½ per cent.

Write for Bulletin 733 for Complete Data

ABRIDGED SPECIFICATIONS

FREQUENCY RANGE—150 to 600 Mc (200 to 50 cm)

OUTPUT POWER—3 to 4 watts up to 400 Mc, decreasing to 1 watt at 600 Mc

OUTPUT IMPEDANCE—approximately 75 ohms

OUTPUT TERMINALS—coaxial jack to facilitate connection by means of a coaxial line to instrument under measurement

CALIBRATION—by means of conversion chart, frequency is determined to ±2½ per cent with no oscillator loading

FREQUENCY CONTROL—slow-motion, lead screw drive with a dial calibrated in division representing approximately 0.01 cm change in wavelength

POWER SUPPLY—filament: 3.65 amperes at [illegible] volts; plate 450 volts (maximum), 50 ma. Th[illegible] Type 757-P1 Power Supply is available to suppl[illegible] fixed filament and plate voltages for operation [illegible] 300 volts from a 115-volt, 60-cycle line

TYPE 757-A U-H-F OSCILLATOR . . $195.00

Type 757-P1 Power Supply 45.00

GENERAL RADIO COMPANY CAMBRIDGE MASSACHUSETTS